Reto B. Keller

Design for Electromagnetic Compatibility - In a Nutshell

Theory and Practice

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Reto B. Keller

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Theory and Practice

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Reto B. Keller
Trachslau, Switzerland

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To my parents Martha and Hans Keller-Camenzind, my wife Nicole Keller-Holdener, and our daughter Valeria Svenja Keller. Thank you for your love and support.

Preface

Electromagnetic compatibility is a topic that is becoming more and more important in our modern and high-tech society. EMC no longer just helps to ensure that the television screen does not flicker when the vacuum cleaner is running. EMC also helps critical applications—such as space missions or medical devices—function reliably. This means that EMC-compliant design receives increased attention, both from engineers and from management.

Design for Electromagnetic Compatibility— In a Nutshell is a book dedicated to development engineers, who design EMC-compliant electrical equipment and electronic devices. This book is to be regarded both as an introduction to EMC for beginners and as a quick reference guide for experts.

The book's content is compressed and focuses on the basics of EMC. Where appropriate, reference is made to further and more advanced literature in the respective chapters. This book contains theoretical and practical knowledge:

Theory. Introduction to EMC, the essential know-how, and the basic concepts of EMC-compliant design

Practice. Discussion of the biggest hurdles to get a product through the EMC test right away and proposal of some universal EMC design guidelines

For years, I have been passionate about EMC. Yes, it sounds odd. I mean: EMC is not a fancy topic, and you cannot talk with your friends about it (at least not with non-engineers), as they get bored quickly. Nevertheless, what is it that makes EMC a fascinating topic for many engineers? For me, EMC means sophisticated theoretical knowledge and practical experience brought together and a responsibility for us—the product development engineers—towards the users and society to release products to the market which are safe and reliable. Enjoy the read!

Einsiedeln, SZ, Switzerland — Reto B. Keller
April 2022

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Relationships are based on four principles: respect, understanding, acceptance and appreciation.

—Mahatma Gandhi

I firmly believe that education is the foundation of prosperity and a free society. Therefore, freely available education is vital in our world. This is why I decided to publish this book open-access. The publication of this book would not have been possible without help. I would like to take this opportunity to express my sincere thanks to all of those who have supported me:

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Contents

1 Introduction 1
1.1 What Is EMC? 1
1.2 EMC vs. EMI 1
1.3 Why Is EMC Important? 2
1.4 EMC Terms and Definitions 2
1.4.1 Emission 3
1.4.2 Immunity 4
1.5 Design for EMC 5
1.6 Summary 6
References 6

2 Regulations and Standards 9
2.1 Big Picture 9
2.2 EMC Compliance 10
2.3 EMC Regulations 10
2.4 EMC Standards 11
2.4.1 What Are EMC Standards? 11
2.4.2 Why Do We Need EMC Standards? 11
2.4.3 Who Writes EMC Standards? 11
2.4.4 EMC Emission Standards 14
2.4.5 EMC Immunity Standards 14
2.4.6 Types of EMC Standards 15
2.4.7 EMC Standards in Different Industries 18
2.4.8 Which EMC Standards to Apply? 19
2.5 Compliance Marks 19
2.6 Global Market Access 21
2.7 Summary 22
References 22

3 Decibel 23
3.1 Gain and Loss [dB] 23
3.2 Absolute Levels [dBm, dBμV, dBμA] 25
3.3 Summary 27
References 30

4 Frequency and Wavelength 33
4.1 EMC and Frequencies 33
4.2 Wavelength vs. Frequency 34
4.2.1 Wavelength in Any Media 34
4.2.2 Wavelength in Insulating Media 35
4.2.3 Wavelength in Vacuum 36
4.2.4 Wavelength in Good Conducting Media 36
4.3 Wavelength of Signals Along Wires, Cables, and PCB Traces 36
4.3.1 Wavelength of Signals Along Blank Wires 37
4.3.2 Wavelength of Signals Along Cables and PCB Traces 37
4.3.3 Summary 39
References 40

5 Time-Domain and Frequency-Domain 41
5.1 Fourier Analysis 41
5.2 Frequency Spectra of Digital Signals 43
5.3 Bandwidth of Digital Signals 45
5.4 Ringing and Frequency Spectrum 46
5.5 Summary 47
References 48

6 RF Parameters 49
6.1 Reflection Coefficient $\underline{\Gamma}$ 49
6.2 Voltage Standing Wave Ratio (VSWR) 50
6.3 Return Loss (RL) 51
6.4 Insertion Loss (IL) 52
6.5 Scattering Parameters 53
6.6 Signal-to-Noise Ratio 55
6.7 Noise Factor and Noise Figure 55
6.8 1 dB Compression Point 56
6.9 Spectrum Analyzer Terms 57
6.9.1 Frequency Range 58
6.9.2 Resolution Bandwidth 58
6.9.3 Video Bandwidth 58
6.9.4 Sweep Time 58
6.9.5 Detectors 61
6.10 Summary 61
References 62

7 Transmission Lines ... 65
7.1 What Is a Transmission Line ... 65
7.2 When to Consider Transmission Lines ... 66
7.2.1 Rule of Thumb for $\mathbf{l_{critical}}$ in the Frequency Domain ... 66
7.2.2 Rule of Thumb for $\mathbf{l_{critical}}$ in the Time Domain ... 67
7.2.3 Critical Length $\mathbf{l_{critical}}$ ... 68
7.3 Characteristic Impedance Z_0 ... 69
7.3.1 Characteristic Impedance Z_0 of Any Transmission Line ... 69
7.3.2 Characteristic Impedance Z_0 of Lossless Transmission Lines ... 70
7.3.3 Characteristic Impedance Z_0 of Parallel Wires ... 71
7.3.4 Characteristic Impedance Z_0 of Twisted Pairs ... 71
7.3.5 Characteristic Impedance Z_0 of Coaxial Cables ... 72
7.3.6 Characteristic Impedance Z_0 of Microstrip Lines ... 73
7.3.7 Characteristic Impedance Z_0 of Coplanar Waveguide with Reference Plane ... 73
7.3.8 Characteristic Impedance Z_0 of Centered Striplines ... 74
7.4 Per-Unit-Length Inductance L' and Capacitance C' ... 75
7.4.1 L' and C' of Two-Wire Lines ... 76
7.4.2 L' and C' of Wires Above Ground Plane ... 77
7.4.3 L' and C' of Coaxial Cables ... 79
7.4.4 L' and C' of PCB Traces ... 80
7.5 Propagation Constant $\boldsymbol{\gamma}$ ... 82
7.6 Input Impedance of Transmission Lines ... 85
7.6.1 Input Impedance of Any Transmission Line ... 85
7.6.2 Input Impedance of a Lossless Transmission Line ... 85
7.6.3 Input Impedance of a Transmission Line At $\mathbf{l = \lambda/2}$ ... 86
7.6.4 Impedance of a Transmission Line At $\mathbf{l = \lambda/4}$ ... 86
7.6.5 Impedance of a Matched Transmission Line ... 86
7.6.6 Impedance of a Shorted Transmission Line ... 87
7.6.7 Impedance of an Open Transmission Line ... 87
7.7 High-Frequency Losses ... 87
7.8 Loss Tangent $\tan(\delta)$ ... 88
7.9 Balanced vs. Unbalanced Transmission Lines ... 89
7.10 Single-Ended vs. Differential Interfaces ... 90
7.11 Summary ... 91
References ... 94

8 Electromagnetic Fields ... 95
8.1 The Electromagnetic Field ... 95
8.2 Electromagnetic Field Characteristics ... 96
8.3 Near-Field vs. Far-Field ... 97
8.3.1 Near-Field ... 97
8.3.2 Far-Field ... 100

8.4 Intrinsic Impedance η 100
8.4.1 Intrinsic Impedance of Any Media 100
8.4.2 Intrinsic Impedance of Magnetic Lossless Media 101
8.4.3 Intrinsic Impedance of Lossless Insulators 101
8.4.4 Intrinsic Impedance of Free-Space 101
8.4.5 Intrinsic Impedance of Good Conductors 102
8.5 Wave Impedance $\mathbf{Z_w}$ 102
8.5.1 Wave Impedance of Any Wave 102
8.5.2 Wave Impedance vs. Distance 103
8.5.3 Wave Impedance in the Near-Field of E-Field Antennas 104
8.5.4 Wave Impedance in the Near-Field of H-Field Antennas 105
8.5.5 Wave Impedance of Plane Waves 106
8.6 Summary 106
References 108

9 Antennas 111
9.1 Antennas and EMC Testing 111
9.2 Isotropic Radiator 112
9.3 Antenna Directivity D 113
9.4 Antenna Gain G 113
9.5 Effective Aperture A_e 116
9.6 Antenna Factor AF 116
9.7 Ground Plane Reflection and Multipath Effect 118
9.8 Intended Antennas 119
9.8.1 E-Field Antennas and Emission Testing 119
9.8.2 E-Field Antennas and Immunity Testing 120
9.8.3 H-Field Antennas and Emission Testing 122
9.8.4 H-Field Antennas and Immunity Testing 123
9.9 Unintended Antennas 125
9.9.1 E-Field from Differential-Mode Currents in Small Loops 125
9.9.2 E-Field from Common-Mode Currents in Short Cables 126
9.9.3 Maximum E-Field from Common-Mode Signals 127
9.9.4 Maximum E-Field from PCB with Attached Cable 130
9.10 Free-Space Path Loss 131
9.11 Link Budget 132
9.12 Summary 133
References 133

10 Skin Effect 135
10.1 Skin Depth 135
10.2 DC vs. AC Resistance 138

10.3 Surface Resistance ... 140
10.4 Summary ... 142
References ... 142

11 Components ... 145
11.1 Conductors ... 145
11.1.1 Definition of a Conductor ... 145
11.1.2 Conductor Equivalent Circuits ... 145
11.1.3 Resistance of a Conductor ... 148
11.1.4 Internal Inductance of a Conductor ... 149
11.1.5 Internal Impedance of a Conductor ... 151
11.1.6 External Inductance of a Single Conductor ... 152
11.1.7 Conductor with Return Current Path ... 153
11.2 Resistors ... 154
11.3 Capacitors ... 157
11.4 Inductors ... 164
11.5 Ferrites Beads ... 169
11.5.1 Cable Mount Ferrite Beads ... 169
11.5.2 PCB Mount Ferrite Beads ... 171
11.6 Common-Mode Chokes ... 173
11.7 Baluns ... 178
11.8 Clamping Devices ... 180
11.8.1 Varistors ... 180
11.8.2 TVS Diodes ... 183
11.9 Crowbar Devices ... 185
11.10 Summary ... 186
References ... 188

12 Noise Coupling ... 189
12.1 Coupling Paths ... 189
12.1.1 Conductive Coupling ... 191
12.1.2 Capacitive Coupling ... 192
12.1.3 Inductive Coupling ... 195
12.1.4 Electromagnetic Coupling ... 197
12.2 Differential-Mode vs. Common-Mode ... 198
12.2.1 Differential-Mode vs. Common-Mode Voltage ... 199
12.2.2 Differential-Mode vs. Common-Mode Current ... 200
12.3 Differential-Mode Noise Sources ... 200
12.3.1 Conductive Coupling to Differential-Mode Noise ... 201
12.3.2 Capacitive Coupling to Differential-Mode Noise ... 201
12.3.3 Inductive Coupling to Differential-Mode Noise ... 202
12.3.4 Common-to-Differential-Mode Conversion ... 202
12.4 Common-Mode Noise Sources ... 203
12.4.1 Capacitive Coupling to Common-Mode Noise ... 204
12.4.2 Electromagnetic Coupling to Common-Mode Noise ... 204

12.4.3 Reference Point Noise to Common-Mode Noise 205
12.4.4 Differential-to-Common-Mode Conversion 205
12.5 Summary 208
References 209

13 Shielding 211
13.1 Shielding Theory 212
13.2 Shielding Effectiveness 212
13.3 Far-Field Shielding 214
13.3.1 Reflection Loss R for Far-Field Shielding 214
13.3.2 Absorption Loss A for Far-Field Shielding 215
13.3.3 Multiple-Reflection Loss Correction M for Far-Field Shielding 216
13.4 Near-Field Shielding 216
13.4.1 Near-Field Shielding of Electric Sources 217
13.4.2 Near-Field Shielding of Magnetic Sources 218
13.4.3 Low-Frequency Magnetic Field Shielding 219
13.5 Slots and Apertures 219
13.5.1 Single Aperture 219
13.5.2 Multiple Apertures 221
13.5.3 Waveguide Below Cutoff 223
13.6 Grounding of Shields 224
13.7 Cable Shields 225
13.7.1 Transfer Impedance $\mathbf{Z}_t$ 225
13.7.2 Cable Shielding Against Capacitive Coupling 225
13.7.3 Cable Shielding Against Inductive Coupling 227
13.7.4 Grounding of Cable Shields 231
13.8 Summary 232
References 233

14 Grounding 235
14.1 Types of Ground 235
14.2 Ground Symbols 235
14.3 EMC Grounding Philosophy 236
14.4 Return Current Path on PCB Ground Planes 237
14.5 Grounding of Systems 238
14.5.1 Single-Point Ground Systems 239
14.5.2 Multipoint Ground Systems 239
14.5.3 Hybrid Ground Systems 240
14.6 Ground Loops 241
14.7 Summary 243
References 244

15 Filtering 245
15.1 Filter Characterization 245
15.1.1 Time-Domain: Step Response 246
15.1.2 Frequency-Domain: Frequency Response 248

15.2 Low-Pass Filters 249
15.3 High-Pass Filters 250
15.4 Band-Pass Filters 250
15.5 Band-Stop Filters 250
15.6 Active and Passive Filters 251
15.7 Differential- and Common-Mode Filters 253
15.7.1 Differential-Mode Filters 254
15.7.2 Common-Mode Filters 254
15.8 Mains Supply Filters 255
15.9 Transient Suppression Filters 256
15.10 Digital Filters 259
15.11 Summary 261
References 262

16 EMC Design Guidelines 265
16.1 Most Common EMC Issues in Practice 265
16.2 Guideline # 1: Never Route Signals Over Split Reference Planes 266
16.3 Guideline # 2: Always Consider the Return Current 266
16.4 Guideline # 3: Decoupling—Use Low-Inductance Capacitors AND Planes 267
16.5 Guideline # 4: Use Ground Planes on PCB for Shielding 267
16.6 Guideline # 5: Route High-Frequency Signals Adjacent to a Plane 268
16.7 Guideline # 6: Control Rise- and Fall-Time 268
16.8 Guideline # 7: Keep Clock lines As Short As Possible 269
16.9 Guideline # 8: Fill Top and Bottom Layers with Circuit GND 269
16.10 Guideline # 9: Add Stitching Vias Around High-Speed Signal Vias 271
16.11 Guideline # 10: Add a Capacitor Close to Every Pin of a Connector 271
16.12 Guideline # 11: Connect Circuit GND to Chassis at IO Area 272
16.13 Guideline # 12: Lay Cables Along Chassis (GND/Earth) 272
16.14 Guideline # 13: Don't Use Cable Shield as Signal Conductor for Low-Frequency Signals 273
16.15 Guideline # 14: Cable Shield Grounding on Only One End for Low-Frequency Signals 273
16.16 Guideline # 15: Cable Shield Grounding on Both Ends for High-Frequency Signals 274
16.17 Guideline # 16: Minimize Loop Area of Signals in Cables 274
16.18 Guideline # 17: Avoid Electrical Balance Changes 274
References 276

A Constants 277

B Complex Numbers 279
B.1 Definition and Notation 279
B.2 Cartesian Complex Plane 279
B.3 Polar Complex Plane 280
B.4 Complex Conjugate 281
B.5 Mathematical Operations with Complex Numbers 281
B.6 Impedances in the Complex Plane 281

C Vectors 283
C.1 Vector Definition 283
C.2 Sum of Vectors 283
C.3 Vector Multiplication with Scalars 284
C.4 Vector Dot Product 284
C.5 Vector Cross Product 285
References 286

D Electromagnetism 287
D.1 Voltage 287
D.2 Current 288
D.3 Impedance 288
D.4 Power 289
D.5 Complex Permittivity 289
D.6 Complex Permeability 290
D.7 Electric Field 291
D.8 Electric Flux 292
D.9 Electric Flux Density 292
D.10 Displacement Current Density 293
D.11 Polarization Density 294
D.12 Electric Current Density 294
D.13 Magnetic Field 295
D.14 Magnetic Flux 295
D.15 Magnetic Flux Density 296
D.16 Electromagnetic Field 297
References 298

E Maxwell's Equations and Lorentz's Force Law 299
E.1 Gauss' Law for Electricity 299
E.2 Gauss' Law for Magnetism 300
E.3 Faraday's Law 300
E.4 Ampére's Law 301
E.5 Lorentz's Force Law 302
Reference 302

F Integral Transforms 303
F.1 Fourier Series 303
F.2 Fourier Transform 304
F.3 Discrete Fourier Transform (DFT) 304
F.4 Discrete-Time Fourier Transform (DTFT) 305
F.5 Laplace Transform 305
F.6 Z-Transform 306
References 306

G Material Properties 307
G.1 Conductor Properties 307
G.2 Insulator Properties 310
G.3 PCB Materials 311
References 313

H Galvanic Series 315
H.1 Corrosion 315
H.2 Corrosion Prevention 316
H.3 Standard Electrode Potential Series 316
H.4 Seawater Galvanic Series 317
References 319

I Triboelectric Effect 321
I.1 How Electrostatic Charges Are Produced 321
I.2 Triboelectric Series 322
References 323

J Country-Specific EMC Regulations 325
J.1 Australia and New Zealand 325
J.2 Brazil 326
J.3 Canada 327
J.4 China 327
J.5 Eurasian Economic Union 328
J.6 European Union 329
J.7 India 331
J.8 Japan 332
J.9 Republic of Korea 333
J.10 Switzerland 333
J.11 Taiwan 334
J.12 Turkey 335
J.13 United Kingdom 336
J.14 United States of America 336
References 339

K EMC Test Setups and Equipment 343
K.1 Pre-compliance vs. Compliance Testing 343
K.2 RF Emission: CISPR 11 344
K.2.1 CISPR 11: Applicability 345
K.2.2 CISPR 11: Group 1, 2 345
K.2.3 CISPR 11: Class A, B 347
K.2.4 CISPR 11: Test Setup 348
K.2.5 CISPR 11: Test Limits 349
K.3 RF Emission: CISPR 32 357
K.3.1 CISPR 32: Applicability 357
K.3.2 CISPR 32: Class A, B 357
K.3.3 CISPR 32: Limits 358
K.4 Harmonic Currents: IEC 61000-3-2 358
K.4.1 IEC 61000-3-2: Applicability 359
K.4.2 IEC 61000-3-2: Class A, B, C, D 361
K.4.3 IEC 61000-3-2: Test Setup 362
K.4.4 IEC 61000-3-2: Limits 363
K.5 Flicker: IEC 61000-3-3 365
K.5.1 IEC 61000-3-3: Applicability 366
K.5.2 IEC 61000-3-3: Test Setup 366
K.5.3 IEC 61000-3-3: Limits 367
K.6 ESD: IEC 61000-4-2 369
K.6.1 IEC 61000-4-2: Test Setup 370
K.6.2 IEC 61000-4-2: Test Levels 373
K.7 RF Radiated Immunity: IEC 61000-4-3 373
K.7.1 IEC 61000-4-3: Test Setup 374
K.7.2 IEC 61000-4-3: Test Levels 376
K.8 EFT: IEC 61000-4-4 377
K.8.1 IEC 61000-4-4: Test Setup 377
K.8.2 IEC 61000-4-4: Test Levels 378
K.9 Surge: IEC 61000-4-5 380
K.9.1 IEC 61000-4-5: Test Setup 380
K.9.2 IEC 61000-4-5: Test Levels 382
K.10 RF Conducted Immunity: IEC 61000-4-6 384
K.10.1 IEC 61000-4-6: Test Setup 384
K.10.2 IEC 61000-4-6: Test Levels 388
K.11 Magnetic Field Immunity: IEC 61000-4-8 388
K.11.1 IEC 61000-4-8: Applicability 389
K.11.2 IEC 61000-4-8: Test Setup 390
K.11.3 IEC 61000-4-8: Test Levels 391
K.12 AC Dips: IEC 61000-4-11 392
K.12.1 IEC 61000-4-11: Applicability 393
K.12.2 IEC 61000-4-11: Test Setup 393
K.12.3 IEC 61000-4-11: Test Levels 393

K.13 Generic Immunity Levels for Residential Environments: IEC 61000-6-1 ... 395
K.14 Generic Immunity Levels for Industrial Environment: IEC 61000-6-2 ... 396
K.15 Generic Emission Limits for Residential Environments: IEC 61000-6-3 ... 397
K.16 Generic Emission Limits for Industrial Environment: IEC 61000-6-4 ... 397
K.17 Generic Emission Limits for Light-Industrial Environment: IEC 61000-6-8 ... 397
References ... 398

L American Wire Gauge (AWG) ... 401

M Rules of Thumb ... 403
M.1 Air Breakdown Voltage ... 403
M.2 Amplitude of Square Wave Harmonics ... 403
M.3 Antenna Input Power for Desired E-Field Strength ... 405
M.4 Bandwidth of Digital Signals ... 405
M.5 Critical Length of Transmission Lines ... 405
M.6 Characteristic Impedance of Wires ... 405
M.7 Capacitance Per-Unit-Length ... 406
M.8 Inductance Per-Unit-Length ... 406
M.9 CMOS Input Impedance ... 406
M.10 E-Field from Differential-Mode Currents in Small Loops ... 407
M.11 E-Field from Common-Mode Currents in Short Cables ... 407
M.12 E-Field from Common-Mode Current with Cable at Resonance ... 407
M.13 E-Field from Common-Mode Voltage with Cable at Resonance ... 408
M.14 E-Field Strength for Given Antenna Power ... 408
M.15 Galvanic Corrosion ... 409
M.16 Return Current on PCB Planes ... 409
M.17 Via Capacitance and Inductance ... 409
References ... 409

Index ... 411

Acronyms

The key test for an acronym is to ask whether it helps or hurts communication.

—Elon Musk

List of the most common *abbreviations* and *acronyms* in the field of electromagnetic compatibility used throughout this book:

AAMI	Association for the Advancement of Medical Instrumentation
AC	Alternating Current
ADC	Analog-to-Digital Converter
AF	Antenna Factor
AM	Amplitude Modulation
ANSI	American National Standards Institute
BCI	Bulk Current Injection
CB	Certified Body
CE	Conducted Emissions
CEN	European Committee for Standardization
CENELEC	European Committee for Electrotechnical Standardization
CI	Conducted Immunity
CIGRE	International Council on Large Electric Systems
CISPR	International Special Committee on Radio Interference
CS	Conducted Susceptibility
CSA	Canadian Standards Association
dB	Decibel, logarithmic unit (dimensionless)
dBd	Gain of an antenna relative to a dipole antenna/radiator
dBi	Gain of an antenna relative to an isotropic antenna/radiator
dBm	Decibel milliwatt, logarithmic unit
dBW	Decibel Watt, logarithmic unit
DC	Direct Current
DFT	Discrete Fourier Transform, mathematical operation
DPI	Direct Power Injection
E3	Electromagnetic Environmental Effects
EEE	Electromagnetic Environmental Effects
EFT	Electrical Fast Transients, also called bursts or EFT/B

EM	Electromagnetic
EMC	Electromagnetic Compatibility
EMI	Electromagnetic Interference
EN	European Standards
ERP	Effective Radiated Power
ESD	Electrostatic Discharge
ESL	Equivalent Series Inductance (of a capacitor)
ESR	Equivalent Series Resistance (of a capacitor)
ETSI	European Telecommunications Standards Institute
EUT	Equipment Under Test
FAC	Fully Anechoic Chamber, type of radiated emissions test site
FAR	Fully Anechoic Room, type of radiated emissions test site
FCC	Federal Communications Commission, USA
FDA	U.S. Food and Drug Administration
FFT	Fast Fourier Transform, fast algorithm for calculating the DFT
GDT	Gas Discharge Tube
GND	Ground
GTEM	Gigahertz Transverse Electromagnetic (Mode)
HBM	Human Body Model
HEMP	High-altitude Electromagnetic Pulse
HF	High-Frequency
HV	High-Voltage
IDM	Imbalance Difference Modeling
IEC	International Electrotechnical Commission
IECEE	International Electrotechnical Commission for Electrical Equipment
IEEE	Institute of Electrical and Electronics Engineers
IL	Insertion Loss, a high-frequency parameter
ISM	Industrial, Scientific, and Medical
ISO	International Organization for Standardization
ITE	Information Technology Equipment
ITU	International Telecommunication Union, specialized agency of the United Nations
LF	Low-Frequency
LISN	Line Impedance Stabilization Network
LV	Low-Voltage
MLCC	Multilayer Ceramic Capacitor
MV	Medium Voltage
NEMP	Nuclear Electromagnetic Pulse
OATS	Open Area Test Site, type of radiated emissions test site
PCB	Printed Circuit Board
PCBA	Printed Circuit Board Assembled
RAM	Radiation-Absorbent Material
RE	Radiated Emissions
RF	Radio-Frequency
RFI	Radio-Frequency Interference

RI	Radiated Immunity
RL	Return Loss, a high-frequency parameter
RMS	Root Mean Square
RoHS	Restriction of (the use of certain) Hazardous Substances in electrical and electronic Equipment
RS	Radiated Susceptibility
RTCA	Radio Technical Commission for Aeronautics, U.S. volunteer organization
SAC	Semi-Anechoic Chamber, type of radiated emissions test site
SAE	Society of Automotive Engineers, US standards developing organization
SMD	Surface Mount Device
SMT	Surface Mount Technology
SNR	Signal-to-Noise Ratio
SPICE	Simulation Program with Integrated Circuit Emphasis
TEM	Transverse Electromagnetic (Mode)
VSWR	Voltage Standing Wave Ratio

Symbols and Units

In science there is only physics; all the rest is stamp collecting.

—Lord Kelvin

Table 1 Symbols, dimensions, and units. Part 1/2

Quantity	Symbol	Definition	Unit
Admittance	Y	$Y=I/V$	Siemens S
Attenuation constant	α		1/m
Capacitance	C	$C=Q/V$	Farad F
Capacitive Reactance	X_C	$X_C=1/(2\pi fC)$	Ohm Ω
Complex dielectric	$\underline{\varepsilon}$	$\underline{\varepsilon}=\varepsilon'-j\varepsilon''$	F/m
Complex permeability	$\underline{\mu}$	$\underline{\mu}=\mu'-j\mu''$	H/m
Complex propagation	$\underline{\gamma}$	$\underline{\gamma}=\alpha+j\beta$	1/m
Conductivity	σ	$\sigma=J/E$	S/m
Current	I	$\lvert I\rvert=dQ/dt$	Ampere A
Current density	J	$\lvert J\rvert=dI/dA$	A/m^2
Dielectric conductivity	σ	$\sigma=\omega\varepsilon''$	S/m
Dielectric constant	ε'	$\varepsilon'=\varepsilon_r'\varepsilon_0=\lvert D\rvert/\lvert E\rvert$	F/m
Dissipation factor, loss factor	D	$D=\tan(\delta)=\varepsilon''/\varepsilon'$	Dimensionless
Electric charge	Q		Coulomb C
Electric field strength	E	$E=F/Q'$	V/m
Electric flux density	D	$\varepsilon_0\varepsilon_r'E$	C/m^2
Electric loss factor	ε''	$\varepsilon''=J_{loss}/(\omega E)$	F/m
Electric polarization	P	$P=D-\varepsilon_0 E$	FV/m^2
Electric susceptibility	χ_e	$\chi_e=\varepsilon_r'-1$	Dimensionless
Electromotive force	emf	$\oint(E\cdot dl)$	Volt V
Force	F	$F=qE+qv\times B$	Newton N
Frequency, angular frequency	f, ω	$f=1/T$, $\omega=2\pi f$	1/sec=Hz, 1/rad
Impedance	Z	$Z=V/I$	Ohm Ω

Table 2 Symbols, dimensions, and units. Part 2/2

Quantity	Symbol	Definition	Unit
Impedance (intrinsic)	η	$\eta = E/H$	Ohm Ω
Inductance	L	$L = V/(dI/dt)$	Henry H
Inductive Reactance	X_L	$X_L = 2\pi fL$	Ohm Ω
Loss tangent (dissipation factor)	$tan(\delta)$	$tan(\delta) = \varepsilon''/\varepsilon'$	Dimensionless
Magnetic field strength	H	$\oint(H \cdot dl) = I$	A/m
Magnetic flux	Φ	$\Phi = B \cdot A$	Weber Wb = Vs
Magnetic flux density	B	$B = \mu_0 (H+M)$	Tesla T
Magnetic loss factor	μ''		H/m
Magnetic permeability	μ'	$\mu' = B/H$	H/m
Magnetic susceptibility	χ_m	$\chi_m = \mu_r' - 1$	Dimensionless
Magnetization	M	$M = B/\mu_0 - H$	A/m
Mutual inductance	M_{ij}	$M_{ij} = \Phi_{ji}/I_j$	H
Phase constant	β	$\beta = 2\pi/\lambda$	1/m
Power	P	$P = dW/dt$	Watt W
Quality factor of capacitor	Q	$Q = 1/tan(\delta)$	Dimensionless
Quality factor of inductor	Q	$Q = 2\pi fL/R$	Dimensionless
Relative dielectic permittivity	ε_r'	$\varepsilon_r' = \varepsilon'/\varepsilon_0$	Dimensionless
Relative magnetic permeability	μ_r'	$\mu_r' = \mu'/\mu_0$	Dimensionless
Resistance	R	$Re(Z)$	Ohm Ω
Resistivity	ρ	$\rho = 1/\sigma$	Ωm
Voltage	V	V=- emf	Volt V
Velocity	v	$v = s/t$	m/sec
Wavelength	λ	$\lambda = 2\pi/\beta$	Meter m

Chapter 1
Introduction

I have not failed. I've just found 10'000 ways that won't work.

—Thomas A. Edison

1.1 What Is EMC?

Electromagnetic compatibility (*EMC*) is an established discipline within the field of electrical and electronics engineering. EMC is the ability of equipment or a system to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to anything in that environment [1]. In other words, each device must have a certain immunity against electromagnetic disturbances, and on the other hand, each device must keep its own electromagnetic emissions low enough to not disturb other devices in its environment (Fig. 1.1).

1.2 EMC vs. EMI

EMI stands for *electromagnetic interference* and is often mixed up with EMC. EMI means that one electronic device A is causing disturbance to another electronic device B, which is in the surrounding of device A.

What is the difference between EMC and EMI? Now, an EMC-compliant product has to be tested on EMI during its development. For an EMC-compliant product, EMI should not happen anymore. This is because EMC-compliant products proved their electromagnetic immunity to be high enough and their electromagnetic emission to be low enough to work seamlessly in its predefined environment.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_1

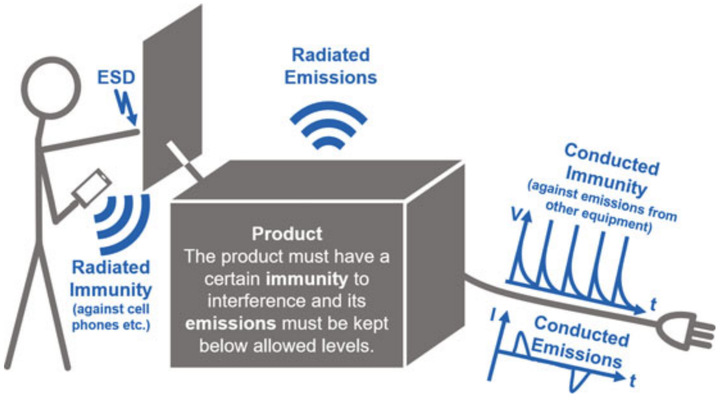

Fig. 1.1 How an EMC design engineer sees a product

1.3 Why Is EMC Important?

EMC-compliant products reduce the risk of undesired interference and disturbances. However, why is that so important? EMC is important because interference and disturbances of electrical equipment can seriously harm people, infrastructure, and the environment. For example, in 1992, a woman died because the heart machine of the ambulance shut down every time the technicians turned on their radio transmitter to ask for advice [12]. Another example is the explosion of the Texaco refinery in Milford Haven UK, on the 24th of July 1994, which was caused by an electrical storm giving rise to power surges which tripped out a number of pump motors while leaving others running. The explosion led to 26 people being sustainably injured and damage of £48 million [12]. These are just two of numerous examples and show us that taking care of EMC-compliant design is not just a necessity for selling products; it means a safer world with reliable products and satisfied customers.

1.4 EMC Terms and Definitions

The discipline of EMC can primarily be divided into the subjects of *emission* and *immunity* (or *susceptibility*). Figure 1.2 shows the most important terms and definitions in EMC. Other EMC abbreviations and acronyms can be found in chapter “Acronyms” in the front matter of this book on page xxiii.

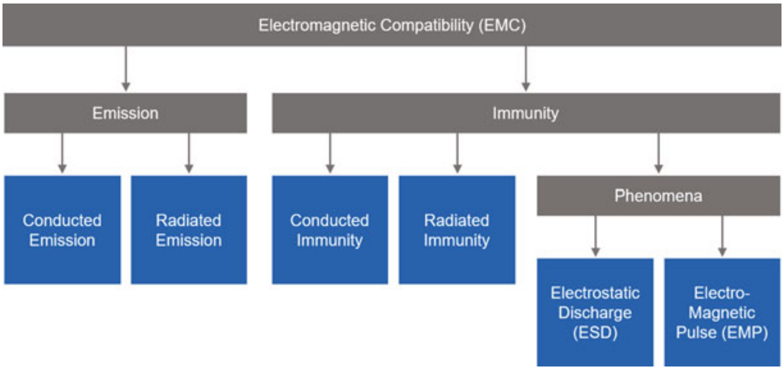

Fig. 1.2 EMC terms and definitions

1.4.1 Emission

It is distinguished between *conducted emissions* (CE) and *radiated emissions* (RE).

- **Conducted emissions.** Conducted emissions are measured at cables that are connected to the equipment under test (EUT). For commercial and industrial electrical/electronic equipment, it is distinguished between different emission aspects:
  - **RF conducted emissions.** Goal: prevent connected cables from radiating and avoid the interference of connected equipment. The frequency range of the RF emission measurements for commercial and industrial products goes from 150 kHz to 30 MHz (CISPR 32 [2], FCC 47 CFR Part 15 [3]). However, the frequency range depends on the industry and may be as wide as from 30 Hz to 40 GHz for defense and military products (MIL-STD-461G [13]).
  - **Harmonics.** Goal: limitation of harmonic currents injected into the public mains supply system. The mains supply input current of many devices and machines is often not a harmonic sine wave.
  - **Flicker.** Goal: limitation of voltage changes, voltage fluctuations, and flicker in public low-voltage supply systems. *Flicker* is the impression of unsteadiness of visual sensation induced by a light stimulus whose luminance or spectral distribution fluctuates with time [1].

- **RF radiated emissions.** Goal: prevent disturbance of nearby electrical and electromechanical equipment. Radiated emissions are measured in an anechoic or semi-anechoic chamber or at an open area test site (OATS). For commercial and industrial products, the frequency range of radiated emission measurements goes from 30 MHz to 6 GHz (CISPR 32 [2]) or 30 MHz up to 40 GHz (FCC 47 CFR Part 15 [3]). However, the frequency range depends on the industry and may be as wide as from 10 kHz to 40 GHz for defense and military products (MIL-STD-461G [13]).

1.4.2 Immunity

It is distinguished between conducted immunity/susceptibility (CI, CS), radiated immunity/susceptibility (RI, RS), and immunity against phenomena like electrostatic discharge (ESD) or electromagnetic pulses (EMP).

- **Conducted immunity.** Conducted immunity tests are performed at cables that are connected to the EUT. For commercial and industrial electrical/electronic equipment, it is distinguished between different immunity aspects:
  - **RF conducted immunity.** Goal: functional immunity to conducted disturbances induced by RF fields. The frequency range of the conducted RF immunity tests for commercial and industrial products goes from 150 kHz to 80 MHz (IEC 61000-4-6 [9]).
  - **EFT.** Goal: functional immunity to repetitive electrical fast transients (EFT, bursts), such as those originating from switching transients (interruption of inductive loads, relay contact bounce, etc.). Test voltages vary from 0.5 kV to 4 kV (IEC 61000-4-4 [8]).
  - **Surge.** Goal: functional immunity to unidirectional surges caused by overvoltages from switching and lightning transients. Test voltages vary from 0.5 kV to 4 kV (IEC 61000-4-5 [11]).
  - **Dips.** Goal: functional immunity to voltage dips, short interruptions, and voltage variations at the power supply ports.
- **Radiated RF field immunity.** Goal: functional immunity to radiated radio-frequency electromagnetic radiation in the far-field (80 MHz to 6 GHz: IEC 61000-4-3 [6]) or in the near-field (26 MHz to 6 GHz: IEC 61000-4-39 [7]).
- **Radiated magnetic field immunity.** Goal: functional immunity to magnetic disturbances at mains power frequencies 50 Hz and 60 Hz (IEC 61000-4-8 [10]) or at wireless charging/inductive power transfer frequencies 9 kHz and 26 MHz (IEC 61000-4-39 [7]).
- **ESD.** Goal: immunity to static electricity discharges, from operators directly and from personnel to adjacent objects. Test voltages vary from 1 kV to 15 kV (IEC 61000-4-2 [5]).
- **EMP.** Goal: protection against electromagnetic pulses (EMP). An EMP is an instantaneous, intense energy field, which can be caused by a nuclear explosion or other pulse-generating devices. If this explosion happens at a high altitude, it is called a HEMP. EMPs are a topic for defense and military applications and not for commercial and industrial products.

1.5 Design for EMC

Design for EMC means considering EMC early during product development, not just at the end of the project. If you test your product for EMC just shortly before the product is to be launched, project delays and budget overruns result. Figure 1.3 shows that the later a defect in the product life cycle is fixed, the more expensive it is. Here are some essential points regarding EMC and product development:

- **EMC concept.** You need to define an EMC concept at the very beginning of the development project (before the first hardware is designed). This EMC concept should especially define the following points:
  - **Grounding.** Define a grounding concept for the product (system grounding), the subsystems of the product (intrasystem grounding), and PCBAs (board-level grounding).
  - **Shielding.** Define if and how to shield sensitive circuits and cables (intrasystem and external cables). Define the bonding of the shields.
  - **Filtering.** Define if and how cables and wires (intrasystem and external cables) have to be filtered. Especially consider ESD, EFT, and surge for cables that leave your product. RF-filtering should be considered for every cable.
- **Iterative testing.** It is recommended to test the EMC performance (emission, immunity) at different product development iterations. There are typically four iterations in hardware development projects: breadboard, prototype, pilot, and series. It is good practice to do pre-compliance testing (testing in-house or at a not fully accredited compliance lab) at the prototype stage and fully compliant EMC testing at the later project stages (pilot, series). Pre-compliance EMC testing is an excellent choice to save time and money. Another advantage of in-house EMC testing is the constant improvement of the EMC knowledge of the development team.

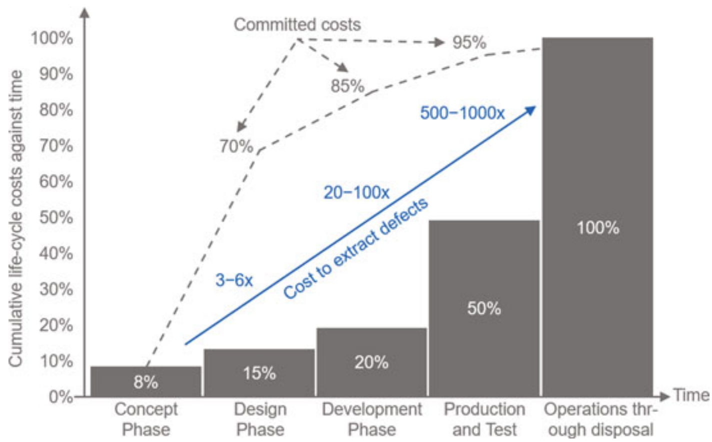

Fig. 1.3 Life cycle costs [4] and the costs of fixing defects at different life cycle stages

1.6 Summary

- **EMC.** EMC is the ability of equipment or a system to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to anything in that environment.
- **EMC tests.** EMC tests can be divided into two categories: emission tests and immunity tests.
- **Coupling paths.** EMI occurs via conduction (connected cables), via radiation (directly from and to a product), or a combination of both.
- **Design for EMC.** Considering EMC design in the early product development stages helps to prevent budget overrun and product launch delays.

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Chapter 2
Regulations and Standards

If you make 10’000 regulations you destroy all respect for the law.

— Winston Churchill

2.1 Big Picture

Products which are sold on the market should have the following properties:

- **Safety.** Be safe.
- **Health.** Do not be health threatening.
- **Environment.** Do not pollute the environment (e.g., RoHS).
- **Quality.** Fulfill expectations regarding quality.
- **Reliability.** Work reliably and as intended in their defined environment (e.g., EMC, temperature, humidity, and altitude).

In order to have only products on the market which fulfill the bullet points above, governments and their legislative bodies issue laws and directives, and products must be regulatory compliant according to this legislation.

A company that is the legal manufacturer of a product must prove compliance with the regulations. This is done with *conformity assessments*. In case a product contains electronics, an EMC conformity assessment is necessary. Such an EMC conformity assessment usually comprises the proof of:

1. The conformity of the product regarding the applicable EMC standards (EMC type testing).
2. The manufacturing process control (quality).

In nearly every country, electronic devices or machines on the market must be EMC compliant, meaning they must fulfill the EMC regulations and standards for the intended use of the products. What EMC regulations and standards are applicable for which product is defined by the country where the product is sold to the end customer.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_2

2.2 EMC Compliance

EMC compliance means that an electronic or electromechanical product is compliant with the laws, directives, and regulations of the country where it is sold to the end customer.

2.3 EMC Regulations

Every government defines its own *EMC regulations* (laws, directives) for its country. However, these national regulations often adopt multinational regulations (e.g., countries in the European Union refer to the EMC directive 2014/30/EU).

The government usually sets up or appoints an organization, commission, or committee responsible for defining the applicable EMC standards (see Fig. 2.1). Such organizations or committees define the applicable EMC standards so that products, which pass the tests defined in the applicable EMC standards, are then compliant with the EMC regulations (laws, directives).

Some authorities (administrations, organizations, commissions, committees) develop the applicable EMC standards for their countries or customs unions themselves. However, in most cases, the authorities adopt the content from the respective international EMC standards into their EMC standards or refer to the respective international EMC standards. Further information about specific countries and their responsible authorities for EMC regulations and standards can be found in Sect. 2.5.

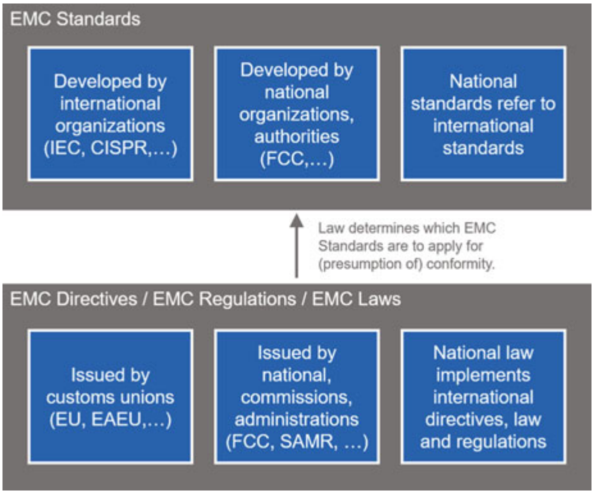

Fig. 2.1 EMC regulations through directives and standards

2.4 EMC Standards

The following chapters focus primarily on international *EMC standards* (IEC, CISPR) and not on local or country-specific standards.

2.4.1 What Are EMC Standards?

EMC standards define terms, rules, and test methods for EMC. Furthermore, they specify limits and minimum test levels for electric and electromagnetic emissions and immunity. Figure 2.2 shows a product and some commonly applied EMC standards for consumer electronic products.

2.4.2 Why Do We Need EMC Standards?

EMC standards help make measurements comparable and repeatable by defining the test methods, test equipment, and test environment. Most importantly, EMC standards aim to bring harmonization to EMC testing, in the best case: global harmonization. Global harmonization of standards reduces trade barriers, and as a most important consequence for society, harmonized EMC standards help increase global prosperity and wealth.

2.4.3 Who Writes EMC Standards?

Standards in EMC are either developed by international, national, or regional organizations and committees on behalf of administrative bodies, or the administrative and/or regulatory bodies word the EMC standards and regulations themselves (e.g., FCC MP-5). Usually, the appointed administrative bodies (e.g., CENELEC for the EU, BSI for the UK, or ACMA for Australia) adopt the international EMC standards written by IEC/CISPR or ISO, word for word. The *International Electrotechnical Commission* (*IEC*) is the most important organization when it comes to EMC standards development. Within the IEC, the following committees have the lead:

- **IEC Technical Committee 77 (TC 77).** The main scope of the TC 77 covers:
  - Immunity and related items, over the whole frequency range: Basic EMC Publications and Generic EMC Standards.
  - Emission in the low-frequency range ($f <= 9\,\text{kHz}$, e.g., harmonics and voltage fluctuations): Basic, Generic, and Product (Family) EMC Standards.

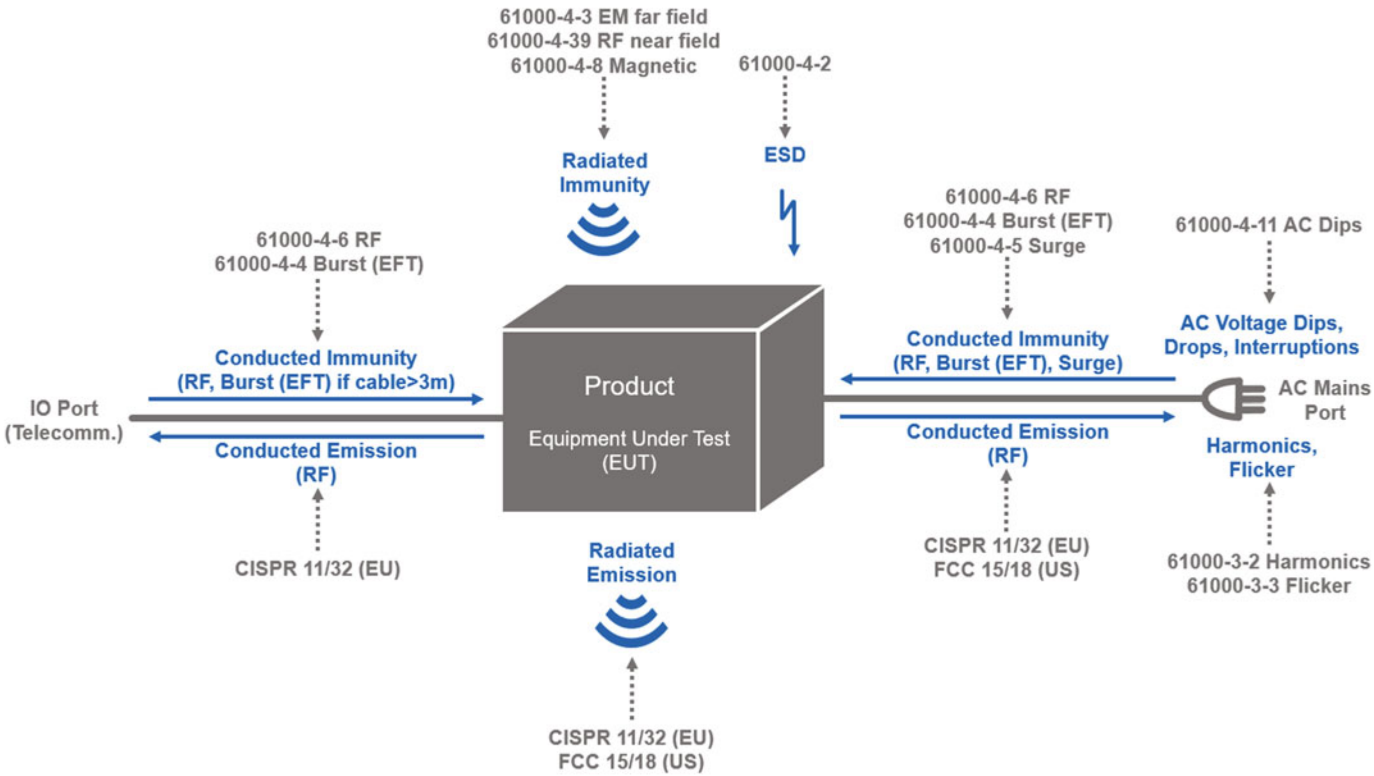

Fig. 2.2 Most commonly applied EMC standards for consumer electronic products with <16 A/phase

  - Emission in the high-frequency range ($f > 9\,\text{kHz}$): disturbances not covered by CISPR 10 (1992), in coordination with CISPR (e.g., mains signaling).

- **CISPR.** The main scope of CISPR is the protection of radio reception in the range 9 kHz to 400 GHz from interference caused by operation of electrical or electronic appliances and systems in the electromagnetic environment.

Here is a (not exhaustive) list of international and national organizations and committees which develop and/or define the applicable EMC standards:

- **International.**
  - IEC Committees for Basic and Generic EMC Standards:
    . Technical Committee 77, Electromagnetic Compatibility (TC 77)
    . International Special Committee on Radio Interference (CISPR)
  - IEC Committees for Product and Product Family EMC Standards.
  - IEC liaison partners:
    . International Organization for Standardization (ISO)
    . ITU Telecommunication Standardization Sector (ITU-T)
    . International Council on Large Electric Systems (CIGRE)
    . The Union of the Electricity Industry (Eurelectric)
    . International Organization of Legal Metrology (OIML)
- **Australia.** Standards Australia (AS).
- **Azerbaijan, Armenia, Belarus, Georgia, Kazakhstan, Kyrgyzstan, Moldova, Russia, Tajikistan, Turkmenistan, Uzbekistan, Ukraine.** Euro-Asian Council for Standardization, Metrology, and Certification (EASC).
- **Canada.** Canadian Standards Association (CSA).
- **China.** Standardization Administration of China (SAC, representing China in national ISO and IEC committees). SAC/TC79: National Radio Interference Standardization Technical Committee (the corresponding Chinese committee to IEC/CISPR). SAC/TC246: National Electromagnetic Compatibility Standardization Technical Committee (the corresponding Chinese committee to IEC/TC77).
- **European Union (EU).**
  - Comité Européen de Normalisation Electrotechniques (CENELEC)
  - European Telecommunications Standards Institute (ETSI)
  - International Council on Large Electric Systems (CIGRE)
  - European Committee for Standardization (CEN)
- **Germany.** The Deutsche Kommission Elektrotechnik Elektronik Informationstechnik (DKE) in DIN and VDE is the organization responsible in Germany for the development of standards, norms, and safety regulations in the fields of electrical engineering, electronics, and information technology.
- **India.** Generally: Bureau of Indian Standards (BIS). Additionally for telecommunication equipment: Telecommunication Engineering Centre (TEC).

- **Japan.**
  - Japanese Industrial Standards Committee (JISC)
  - Japanese Standards Association (JAS)
- **Korea.** Korean Standards Association (KSA).
- **New Zealand.** Standards New Zealand (NZS).
- **Russia.** Federal Agency for Technical Regulation and Metrology (GOST-R).
- **Singapore.** Info-Communications Media Development Authority (IMDA).
- **Turkey.** Turkish Standards Institution (TSE).
- **United Kingdom (UK).** British Standards Institution (BSI).
- **United States (USA).**
  - Association for the Advancement of Medical Instrumentation (AAMI)
  - American National Standards Institute (ANSI)
  - Department of Defense (DoD)
  - Federal Communications Commission (FCC)
  - Radio Technical Commission for Aeronautics (RTCA)
  - Society of Automotive Engineers (SAE)

2.4.4 EMC Emission Standards

EMC emission standards define test methods and emission limits for conducted and radiated electromagnetic emissions. Examples of EMC emission standards for residential, commercial, and industrial products are:

- IEC 61000-3-2—Test methods for harmonic currents
- IEC 61000-3-3—Test methods for flicker
- IEC 61000-6-3—Emission limits for equipment in residential environments
- IEC 61000-6-4—Emission limits for equipment in industrial environments
- IEC 61000-6-8—Emission limits for commercial and light-industrial locations
- CISPR 11—RF emission limits and methods for ISM equipment
- CISPR 14-1—RF emission limits and methods for household appliances
- CISPR 32—RF emission limits and methods for multimedia equipment
- FCC Part 15—RF radiated and conducted emissions (RF devices)
- FCC Part 18—RF radiated and conducted emissions (ISM equipment)

2.4.5 EMC Immunity Standards

EMC immunity standards define test methods and immunity test levels for conducted and radiated electromagnetic disturbances. Examples of EMC immunity standards for residential, commercial, and industrial products are:

- IEC 61000-4-2—Test methods for ESD
- IEC 61000-4-3—Test methods for RF radiated, far-field
- IEC 61000-4-4—Test method for burst (EFT) immunity
- IEC 61000-4-5—Test method for surge immunity
- IEC 61000-4-6—Test methods for RF conducted
- IEC 61000-4-8—Test methods for magnetic field immunity
- IEC 61000-4-11—Test methods for AC dips
- IEC 61000-4-39—Test methods for RF radiated, near-field
- IEC 61000-6-1—Immunity test levels for residential and light-industrial
- IEC 61000-6-2—Immunity test levels for industrial environments
- CISPR 14-2—Immunity test levels for household appliances and electric tools
- CISPR 35—Immunity test levels for multimedia equipment

Performance criteria are used to evaluate the immunity characteristics of the equipment under test (EUT) in the course of an EMC immunity test. In other words, performance criteria describe the loss of function or degradation of performance of the EUT. The pass/fail criteria of an immunity test case are specified in the EMC test plan before testing the EUT. These criteria are typically set to A, B, C, or D. The criteria A, B, C, and D are described in detail in the respective EMC immunity standard [3]:

- **Performance criterion A.** Normal performance within limits specified by the manufacturer, requestor, or purchaser.
- **Performance criterion B.** Temporary loss of function or degradation of performance which ceases after the disturbance ceases and from which the equipment under test recovers its normal performance, without operator intervention.
- **Performance criterion C.** Temporary loss of function or degradation of performance, the correction of which requires operator intervention.
- **Performance criterion D.** Loss of function or degradation of performance which is not recoverable, owing to damage to hardware or software or loss of data.

2.4.6 Types of EMC Standards

The following are classes or types of EMC standards (see Fig. 2.3):

- Basic EMC Publications
- EMC Product Standards
- EMC Product Family Standards
- Generic EMC Standards

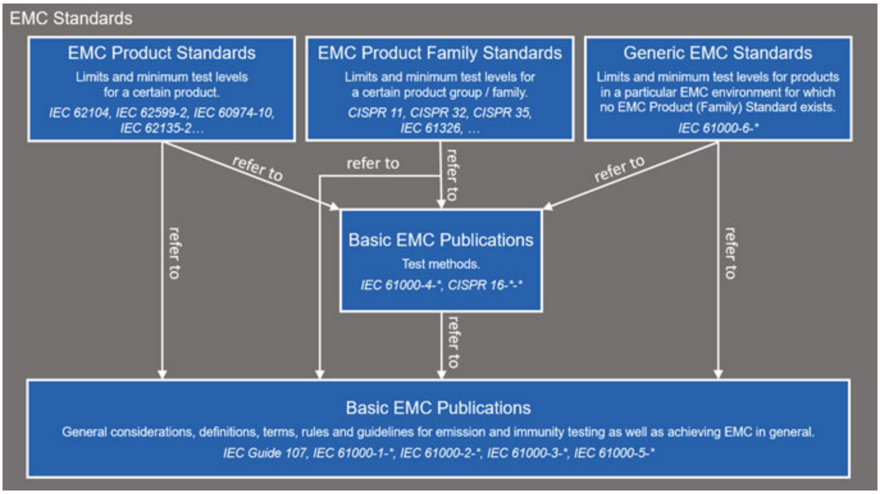

Fig. 2.3 Types of EMC standards and their dependencies

2.4.6.1 Basic EMC Publications

Basic EMC Publications specify terms and conditions for EMC testing, define the rules necessary for achieving electromagnetic compatibility, specify test methods (testing techniques, test setup, test equipment, and environment), and so on. Basic EMC Publications are the EMC standards to which other EMC standards (EMC Product Standards, Generic EMC Standards, etc.) refer.

The IEC Basic EMC Publications are structured into the following categories [1]:

- **General.** Guidance on how to draft an EMC publication, definition of the EMC terminology and vocabulary, general considerations. Examples are IEC Guide 107, IEC 60050-161, IEC TR 61000-1-1, IEC 61000-1-xx, etc.
- **Environment.** Classification and description of different electromagnetic environments and compatibility levels. Examples are IEC TS 61000-2-5, IEC TR 61000-2-3, IEC 61000-2-xx, etc.
- **Emission.** Definition of test setups, testing techniques, test equipment, test environment, and other considerations regarding EMC emission testing and measurement. Examples are IEC 61000-3-2, IEC 61000-3-3, IEC 61000-3-xx, CISPR 16, etc. CISPR 16 is itself a series of publications specifying equipment and methods for measuring disturbances and immunity impacting them at frequencies above 9 kHz.
- **Immunity.** Definition of test setups, testing techniques, test equipment, test environment, and other considerations regarding EMC immunity testing. Examples are IEC 61000-4-1, IEC 61000-4-2, IEC 61000-4-3, IEC 61000-4-xx, etc.

- **Installation/Mitigation.** Installation and mitigation guidelines regarding earthing and cabling, mitigation of external electromagnetic influences, HEMP protection concepts, and so on. Examples are IEC TR 61000-5-1, IEC TR 61000-5-2, IEC 61000-5-xx, etc.

2.4.6.2 EMC Product Standards

EMC Product Standards apply to particular products. EMC Product Standards refer to the relevant Basic EMC Publications (for that particular product) and specify the limits of emission and immunity (the minimum test levels).

Examples of EMC Product Standards are IEC 62104 (DAB receivers), IEC 61851-21 (electric road vehicles charging system), IEC 62599-2 (alarm end electronic security systems), etc.

2.4.6.3 EMC Product Family Standards

EMC Product Family Standards apply to a group of products with common general characteristics and may operate in the same environment and have neighboring fields of application. EMC Product Family Standards refer to the relevant Basic EMC Publications (for that particular product) and specify the limits of emission and immunity (the minimum test levels).

Examples of EMC Product Family Standards are IEC 61967 (integrated circuits), IEC 61326 (electrical equipment for measurement, control, and laboratory use), and IEC 60947 (switchgear and controlgear).

2.4.6.4 Generic EMC Standards

The *Generic EMC Standards* are for products operating in a particular EMC environment (residential/industrial), where a specific EMC Product (Family) Standard does not exist. They are general and somewhat simplified EMC Product Standards while referring to Basic EMC Publications for detailed measurement and test methods. Generic EMC Standards specify a limited number of emission and immunity tests and minimum test levels.

Examples of Generic EMC Standards are IEC 61000-6-1 (immunity standard for residential, commercial, and light-industrial environments), IEC 61000-6-2 (immunity standard for industrial environments), IEC 61000-6-3 (emission standard for equipment in residential environments), IEC 61000-6-4 (emission standard for industrial environments), and IEC 61000-6-8 (emission standard for professional equipment in commercial and light-industrial locations).

2.4.7 EMC Standards in Different Industries

Some industries require their own industry-specific EMC standards. Sometimes they are also called EMC Product (Family) Standards. Table 2.1 lists some popular international and industry-specific EMC standards, e.g., automotive, aviation, lightning, or medical.

Table 2.1 EMC product (Family) standards and industry-specific EMC standard

EMC Product (Family) Standards			
	Emission	**Immunity**	**Remark**
Automotive: Vehicles, **boats** and internal combustion engines	CISPR 12 CISPR 25 CISPR 36	–	CISPR 36: Electric and hybrid electric road vehicles.
Automotive: Road vehicles - Vehicle test methods	–	ISO 11451 ISO 10605	Vehicles.
Automotive: Road vehicles - Component test methods	–	ISO 11452 ISO 7637 ISO 10605	Components.
Avionics: Airborne equipment, aircraft, helicopters	RTCA DO-160 EUROCAE ED-14		RTCA DO-160 and EUROCAE ED-14 are identically worded.
Electricity metering equipment (AC)	–	IEC 62052-11	
Household: Household appliances, electric tools and similar apparatus	CISPR 14-1	CISPR 14-2	
Information technology equipment: ITE, notebooks, telephones	CISPR 32	CISPR 35	CISPR 32 replaced CISPR 22 in 2017. CISPR 35 replases CISPR 24 in 2022.
Lightning: Electrical lighting and similar equipment	CISPR 15	IEC 61547	
Medical equipment: Medical electrical equipment - General requirements for basic safety and essential performance - Collateral Standard: Electromagnetic disturbances - Requirements and tests	IEC 60601-1-2		
Laboratory equipment: Electrical equipment for measurement, control and laboratory use - EMC requirements	IEC 61326-1	IEC 61326-1	
Medical IVD equipment: Electrical equipment for measurement, control and laboratory use - EMC requirements	IEC 61326-2-6		
Medical safety equipment: Electrical equipment for measurement, control and laboratory use - Immunity requirements for safety-related systems and equipment	–	IEC 61326-3-1	
Power unitities: Communication networks and systems for power utility automation	IEC 61850-3		
Relays: Measuring relays and protection equipment	IEC 60255-26	IEC 60255-26	
Residual current operated circuit-breakers with integral overcurrentprotection for household and similar uses	IEC 61009-1		
Ships with a metallic hull: Electrical and electronic installations in ships - Electromagnetic compatibility (EMC)	IEC 60533	IEC 60533	

2.4.8 *Which EMC Standards to Apply?*

As mentioned above in Sect. 2.3, the applicable EMC standards are defined by the responsible governmental administrations, organizations, commissions, or committees. Therefore, the process of finding the applicable standards often differs from country to country. As an example, Fig. 2.4 shows a flow chart based on the EMCD Guide published by the European Commission [4].

Other useful tips:

- Contact the national authorities—where the product will be sold—for legal advice.
- Check which EMC standards your competitors apply.
- Ask your EMC test laboratory for advice on which EMC standards to apply.

2.5 Compliance Marks

Which compliance marks to go for? To decide if and which *compliance marks* you need for your product depends on which countries or customs unions (e.g., EU) you want to sell your product to.

Let's take a step back to the big picture of conformity marks: mandatory marks vs. nonmandatory marks.

1. **Mandatory conformity marks and labels.** Mandatory marks and labels are legally binding and have to be attached to the product. Examples of mandatory marks and labels are:

   - **Australia and New Zealand.** RCM mark (Regulatory Compliance Mark).
   - **Brazil.** ANATEL label (Agência Nacional de Telecomunicações).
   - **Canada.** ISED label (Innovation, Science, and Economic Development).
   - **China.** CCC mark (China Compulsory Certificate).
   - **EEU (Russia).** EAC mark (Eurasian Conformity).
   - **EU.** CE mark (Conformité Européenne).
   - **Japan.** PSE mark (Product Safety Electrical Appliance and Material).
   - **Republic of Korea (South Korea).** KC mark (Korean Certification).
   - **Singapore.** IMDA label (Infocomm Media Development Authority).
   - **Switzerland.** CH mark (Swiss Conformity Mark).
   - **Taiwan.** BSMI mark (Bureau of Standards, Metrology, and Inspection).
   - **UK.** UKCA (United Kingdom Conformity Assessed).
   - **USA.** FCC mark (Federal Communications Commission).

2. **Nonmandatory conformity marks and labels.** Nonmandatory marks are not legally binding and are therefore optional. Examples of nonmandatory marks are:

   - **Canada.** CSA (Canadian Standards Association).

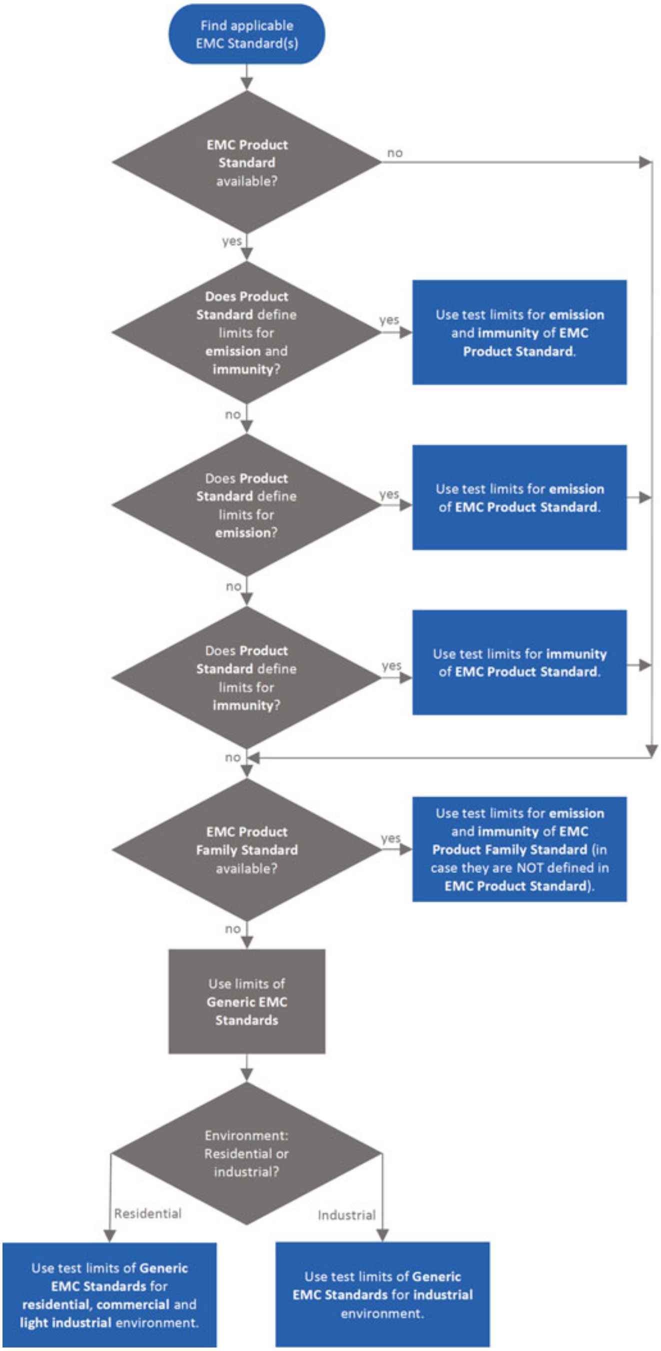

Fig. 2.4 Flow chart based on the EMCD Guide published by the European Commission [4]

- **Germany.** GS (Geprüfte Sicherheit).
- **Japan.** VCCI (Voluntary Control Council for Interference).
- **USA).** UL (Underwriters Laboratory).

More country-specific EMC regulations and requirements can be found in the Appendix in Chap. J.

2.6 Global Market Access

Section 2.5 and Appendix J show that the situation regarding EMC regulations and standards can be confusing. However, there is a great way to get access to the global market with a single EMC test report: test your product according to the International Electrotechnical Commission for Electrical Equipment (IECEE) Certified Body (CB) Scheme. The IECEE CB Scheme is a system for mutual recognition of safety and EMC test certificates of conformity for over 50 countries. It is also a tool for accessing global markets directly when national authorities and regulators, retailers, buyers, and vendors accept CB test certificates and the associate test reports. The same concept of the IECEE CB Scheme does also apply for safety regulations and standards. Figure 2.5 shows the organizational hierarchy of the Member Body (there is one MB per country), the National Certification Body (NCB issues the CB test certificate), and the Certified Body Test Laboratories (CBTLs do perform the actual EMC testing).

This is how the EMC testing for global market access works:

1. **EMC testing.** Find a Certified Body EMC Test Lab (CBTL) and test your product according to the CB Scheme at this CBTL and obtain a CB EMC Test Report for your product. CBTLs are listed on the IECEE website [2].

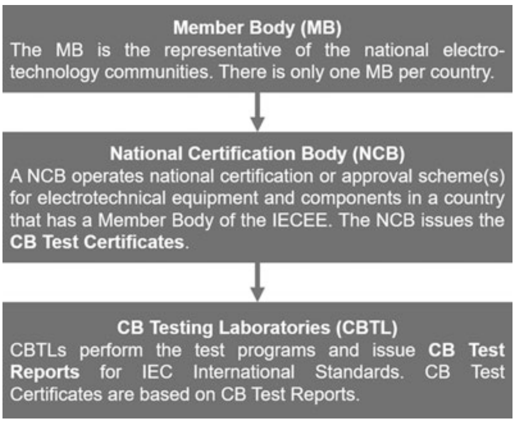

Fig. 2.5 Involved institutions for EMC testing for global market access

2. **CB test certificate.** Request a CB EMC Test Certificate at a National Certification Body (NCB) for your product. NCBs are listed on the IECEE website [2].
3. **Request market access.** To obtain a national EMC certification for a participating CB Scheme country without additional retesting of your product, you have to submit your CB EMC Test Certificate and CB EMC Test Report to the Notified Certified Body (NCB) of that country where you would like to get market access to. Countries that accept the CB Scheme are listed on the IECEE website [2].

2.7 Summary

EMC compliance. EMC compliance means that a product is compliant to the laws, directives, and regulations of the country where it is sold to the end customer.

EMC standards. Each country or customs union defines the applicable EMC standards itself.

Global market access. Testing a product according to the CB Scheme helps to get global market access (for over 50 countries) with minimum test effort.

References

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3. *Electromagnetic compatibility (EMC) – Part 4–5: Testing and measurement techniques – Surge immunity test*. International Electrotechnical Commission (IEC). 2014.
4. *Guide for the EMCD (Directive 2014/30/EU)*. European Commission (EC). Dec. 19, 2018.

Chapter 3
Decibel

If you cannot measure it, you cannot improve it.

—Lord Kelvin

3.1 Gain and Loss [dB]

Let's have a look at the amplifier or attenuation network in Fig. 3.1. The power, voltage, and current gain of this network can be expressed in [dB] as [2]:

$$\text{Power gain in [dB]} = 10 \cdot \log_{10}\left(\frac{P_{out}}{P_{in}}\right) = 10 \cdot \log_{10}\left[\left(\frac{V_{out}}{V_{in}}\right)^2 \cdot \left(\frac{R_{in}}{R_{load}}\right)\right] \tag{3.1}$$

$$\text{Voltage gain in [dB]} = 20 \cdot \log_{10}\left(\frac{V_{out}}{V_{in}}\right) + 10 \cdot \log_{10}\left(\frac{R_{in}}{R_{load}}\right) \tag{3.2}$$

$$\text{Current gain in [dB]} = 20 \cdot \log_{10}\left(\frac{I_{out}}{I_{in}}\right) + 10 \cdot \log_{10}\left(\frac{R_{load}}{R_{in}}\right) \tag{3.3}$$

In case R_{in} and R_{load} are equal (typically 50 Ω), then the following term is equal to zero:

$$10 \cdot \log_{10}\left(\frac{R_{in}}{R_{load}}\right) = 10 \cdot \log_{10}\left(\frac{R_{load}}{R_{in}}\right) = 10 \cdot \log_{10}(1) = 0$$

Now we can write the following for power/voltage/current gain:

$$\textbf{Power gain in [dB]} = 10 \cdot \log_{10}\left(\frac{P_{out}}{P_{in}}\right) \tag{3.4}$$

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_3

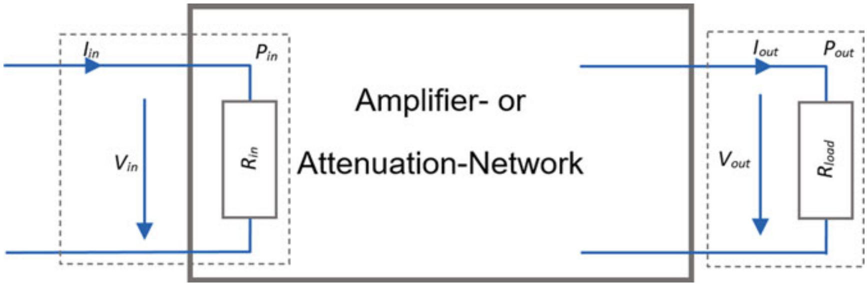

Fig. 3.1 An arbitrary amplifier or attenuation network

$$\textbf{Voltage gain in [dB]} = 20 \cdot \log_{10}\left(\frac{V_{out}}{V_{in}}\right) \tag{3.5}$$

$$\textbf{Current gain in [dB]} = 20 \cdot \log_{10}\left(\frac{I_{out}}{I_{in}}\right) \tag{3.6}$$

Points to remember when it comes to gain and loss calculations in *decibel*:

- **Absolute vs. relative**. Decibels are always ratios of numbers, never an absolute quantity, even if they are named *absolute levels*.
- **Amplification.** If P_{out} is bigger than P_{in}, the gain value in [dB] is positive. This means that in case of an amplification, the power gain in [dB] is positive.
- **Attenuation.** If P_{out} is smaller than P_{in}, the gain value in [dB] is negative. This means that in case of an attenuation (loss), the power gain in [dB] is negative.
- **Gain = −Loss.** Power loss is indicated by a negative decibel power gain. For example, if an interconnection shows a loss of 1 dB, the power gain of that interconnection is −1 dB.
- **Cutoff frequency f_c.** At the cutoff frequency, the output power (P_{out}) is half the input power (P_{in}), and the power/voltage/current gains are all −3 dB.

$$\text{Power gain in [dB] at } f_c = 10 \cdot \log_{10}\left(\frac{1}{2}\right) = -3\text{dB}$$

$$\text{Voltage gain in [dB] at } f_c = 20 \cdot \log_{10}\left(\frac{1}{\sqrt{2}}\right) = -3\text{dB}$$

$$\text{Current gain in [dB] at } f_c = 20 \cdot \log_{10}\left(\frac{1}{\sqrt{2}}\right) = -3\text{dB}$$

- **Ratio to [dB].** Table 3.2 presents some common ratio to [dB] value conversions. For example, if power increases by factor 2, the power/voltage/current gain increases by 3 dB.

Unit	Type Unit	Reference Value	Typical Usage
dBm	Power	1 mW	Wireless and RF applications
dBw	Power	1 W	Wireless and RF applications
dBμV	Voltage	1 μV	Conducted emission
dBμA	Current	1 μA	Conducted emission
dBμV/m	Electric Field Strength	1 μV/m	Radiated emission
dBμA/m	Magnetic Field Strength	1 μA/m	Radiated emission

Fig. 3.2 Commonly used absolute decibel levels in EMC

3.2 Absolute Levels [dBm, dBμV, dBμA]

The most common absolute power, voltage, and current levels in EMC are [dBm], [dBμV], and [dBμA] (Fig. 3.2). They are calculated like this [2]:

$$\textbf{Absolute power level in [dBm]} = 10 \cdot \log_{10}\left(\frac{P}{1\,\mathrm{mW}}\right) \tag{3.7}$$

$$\textbf{Absolute voltage level in [dB}\boldsymbol{\mu}\textbf{V]} = 20 \cdot \log_{10}\left(\frac{V}{1\,\mu\mathrm{V}}\right) \tag{3.8}$$

$$\textbf{Absolute current level in [dB}\boldsymbol{\mu}\textbf{A]} = 20 \cdot \log_{10}\left(\frac{I}{1\,\mu\mathrm{A}}\right) \tag{3.9}$$

More information about physical quantities and their units (also in decibel) can be found in the Appendix D. The same concept of absolute levels can also be applied to electrical fields E [V/m], magnetic fields H [A/m], or radiated power density S [W/m^2]:

$$\textbf{Electric field strengnth in [dB}\boldsymbol{\mu}\textbf{V/m]} = 20 \cdot \log_{10}\left(\frac{E}{1\,\mu\mathrm{V/m}}\right) \tag{3.10}$$

$$\textbf{Magnetic field strength in [dB}\boldsymbol{\mu}\textbf{A/m]} = 20 \cdot \log_{10}\left(\frac{H}{1\,\mu\mathrm{A/m}}\right) \tag{3.11}$$

$$\textbf{Radiated power density in [dBmW/m}^{\mathbf{2}}\textbf{]} = 10 \cdot \log_{10}\left(\frac{S}{1\,\mathrm{mW/m^2}}\right) \tag{3.12}$$

Points to remember when it comes to calculations with absolute power levels in decibel:

- **Zero values.**
  - $0\,\text{dBm} \stackrel{\text{def}}{=} 1\,\text{mW}$
  - $0\,\text{dB}\mu\text{V} \stackrel{\text{def}}{=} 1\,\mu\text{V}$
  - $0\,\text{dB}\mu\text{A} \stackrel{\text{def}}{=} 1\,\mu\text{A}$
  - $0\,\text{dB}\mu\text{V/m} \stackrel{\text{def}}{=} 1\,\mu\text{V/m}$
  - $0\,\text{dB}\mu\text{A/m} \stackrel{\text{def}}{=} 1\,\mu\text{A/m}$
- **Negative values.** A negative [dBm]-value means that the power is smaller than 1 mW. A negative [dBμV]-value means that the voltage is smaller than 1 μV. A negative [dBμA]-value means that the current is smaller than 1 μA.
- **Positive values.** A [dBm]-value bigger than 0 means that the power is higher than 1 mW. A [dBμV]-value bigger than 0 means that the voltage is higher than 1 μV. A [dBμA]-value bigger than 0 means that the current is higher than 1 μA.
- **Gain and loss.** Gain values in [dB] G_{dB} can just be added to the absolute power levels in order to get the output power. The linear calculation with input power P_{in}, linear gain G, and output power P_{out} can be written as:

$$P_{out} = P_{in} \cdot G$$

  In decibel, the output power $P_{out,dB}$ is the sum of the input power $P_{in,dB}$ and the gain G_{dB}:

$$P_{out,dB} = 10 \cdot \log_{10}(P_{out}) = 10 \cdot \log_{10}(P_{in}) + 10 \cdot \log_{10}(G) = P_{in,dB} + G_{dB}$$

  Let's assume a signal with $P_{in,dBm} = 0\,\text{dBm}$ at the input of an amplifier with gain $G_{dB} = 20\,\text{dB}$. The output power is:

$$P_{out,dBm} = P_{in,dBm} + G_{dB} = 0\,\text{dBm} + 20\,\text{dB} = 20\,\text{dBm}$$

- **Never sum up absolute levels.** Do never sum up absolute decibel levels, because adding decibels means multiplying the linear values and therefore:

$$0\,\text{dBm} + 0\,\text{dBm} \stackrel{\text{def}}{=} 1\,\text{mW} \cdot 1\,\text{mW} = 1\,\text{mW}^2 = ?$$

  What does power squared mean? Thus, never add up absolute decibel levels.
- **Sum of absolute levels and decibel.** It is allowed and useful to sum up [dBm]-, [dBμV]-, or [dBμA]-values with gain G values in [dB]:

$$P_{out,lin} = P_{in,lin} \cdot G_{lin} \stackrel{\text{def}}{=} P_{out,dBm} = P_{in,dBm} + G_{dB} = -3\,\text{dBm} + 20\,\text{dB} = 17\,\text{dBm}$$

$$V_{out,lin} = V_{in,lin} \cdot V_{lin} \stackrel{\text{def}}{=} V_{out,dBV} = V_{in,dBV} + G_{dB} = -3\,\text{dBV} + 20\,\text{dB} = 17\,\text{dBV}$$

- **Subtraction of two absolute levels.** Subtracting two absolute [dBm]-, [dBµV]-, or [dBµA]-values is equivalent to computing the ratio of their linear values:

$$\frac{P_{out,lin}}{P_{in,lin}} = \frac{100\,\text{mW}}{1\,\text{mW}} = 100 \stackrel{\text{def}}{=} P_{out,dBm} - P_{in,dBm} = 20\,\text{dBm} - 0\,\text{dBm} = 20\,\text{dB}$$

$$\frac{V_{out,lin}}{V_{in,lin}} = \frac{10\,\text{V}}{1\,\text{V}} = 10 \stackrel{\text{def}}{=} V_{out,dBV} - V_{in,dBV} = 20\,\text{dBV} - 0\,\text{dBV} = 20\,\text{dB}$$

3.3 Summary

The Tables 3.1, 3.2, 3.3 and 3.4 present conversions between different absolute decibel levels, linear rations to relative decibel values and vice versa.

Table 3.1 Conversion between [dBµV], [dBµA], and [dBm] for systems with system impedance $Z_0 = 50\,\Omega$[1]

[dB] Unit Conversion for 50Ω-Systems - [dBm] to [dBµV] to [dBµV]		
dBm [dBm=dBµV-107] [dBm=dBµA-73]	dBµV [dBµV=dBm+107] [dBµV=dB A+34]	dBµA [dBµA=dBm+73] [dBµA=dB V-34]
50	157	123
40	147	113
30	137	103
20	127	93
10	117	83
0	107	73
10	97	63
20	87	53
30	77	43
40	67	33
50	57	23
60	47	13
70	37	3
80	27	7
90	17	17
100	7	27

Table 3.2 Conversion from (linear) ratios to [dB] values and vice versa

Unit Convertion - Voltage/Current/Power Ratio [1] to [dB]						
Voltage Ratio Current Ratio	Power Ratio	[dB]		Voltage Ratio Current Ratio	Power Ratio	[dB]
1	1	0		1	1	0
1.02	1.04	0.17		0.98	0.96	−0.17
1.04	1.08	0.34		0.96	0.92	−0.34
1.07	1.14	0.59		0.93	0.87	−0.59
1.10	1.21	0.83		0.91	0.83	−0.83
1.15	1.32	1.21		0.87	0.76	−1.21
1.20	1.44	1.58		0.83	0.69	−1.58
1.30	1.69	2.28		0.77	0.59	−2.28
1.41	2	3.01		0.71	0.50	−3.01
1.50	2.25	3.52		0.67	0.44	−3.52
1.60	2.56	4.08		0.63	0.39	−4.08
1.73	3	4.77		0.58	0.33	−4.77
1.80	3.24	5.11		0.56	0.31	−5.11
1.90	3.61	5.58		0.53	0.28	−5.58
2	4	6.02		0.50	0.25	−6.02
2.20	4.84	6.85		0.45	0.21	−6.85
2.40	5.76	7.60		0.42	0.17	−7.60
2.50	6.25	7.96		0.40	0.16	−7.96
2.60	6.76	8.30		0.38	0.15	−8.30
2.80	7.84	8.94		0.36	0.13	−8.94
3	9	9.54		0.33	0.11	−9.54
3.25	10.6	10.2		0.31	0.095	−10.2
3.50	12.3	10.9		0.29	0.082	−10.9
3.75	14.1	11.5		0.27	0.071	−11.5
4	16	12.0		0.25	0.063	−12.0
4.50	20.3	13.1		0.22	0.049	−13.1
5	25	14.0		0.20	0.040	−14.0
5.50	30.3	14.8		0.18	0.033	−14.8
6	36	15.6		0.17	0.028	−15.6
6.50	42.3	16.3		0.15	0.024	−16.3
7	49	16.9		0.14	0.020	−16.9
7.50	56.3	17.5		0.13	0.018	−17.5
8	64	18.1		0.13	0.016	−18.1
9	81	19.1		0.11	0.012	−19.1
10	100	20		0.10	0.010	−20
30	900	29.5		0.03	1.11E-03	−29.5
100	1.00E+04	40		0.01	1.00E-04	−40
300	9.00E+04	49.5		0.003	1.11E-05	−49.5
1000	1.00E+06	60		0.001	1.00E-06	−60
3000	9.00E+06	69.5		0.0003	1.11E-07	−69.5
1.00.E+04	1.00E+08	80		1.00E-04	1.00E-08	−80
3.00.E+04	9.00E+08	89.5		3.33E-05	1.11E-09	−89.5
1.00.E+05	1.00E+10	100		1.00E-05	1.00E-10	−100
1.00.E+06	1.00E+12	120		1.00E-06	1.00E-12	−120
1.00.E+07	1.00E+14	140		1.00E-07	1.00E-14	−140
1.00.E+08	1.00E+16	160		1.00E-08	1.00E-16	−160
1.00.E+09	1.00E+18	180		1.00E-09	1.00E-18	−180
1.00.E+10	1.00E+20	200		1.00E-10	1.00E-20	−200

Table 3.3 Conversion between voltages in [V] and [dBμV], between currents in [A] and [dBμA], and between power in [dBm] and [mW] for different system impedances Z_0 [1]

[dB] Unit Conversion - [V], [dBV], [dBμV], [A], [dBA], [dBμA], [dBm]		
To	Calculation	Remark
V	$[\mathbf{V}] = 10^{\left(\frac{[\mathbf{dBV}]}{20}\right)}$	
V	$[\mathbf{V}] = 10^{\left(\frac{([\mathbf{dB\mu V}]-120)}{20}\right)}$	
dBV	$[\mathbf{dBV}] = 20\log_{10}(\mathbf{V})$	
dBV	$[\mathbf{dBV}] = [\mathbf{dB\mu V}] - 120$	
dBμV	$[\mathbf{dB\mu V}] = 20\log_{10}(\mathbf{V}) + 120$	
dBμV	$[\mathbf{dB\mu V}] = [\mathbf{dBm}] + 10\log_{10}(Z) + 90$	Z = system impedance
dBμV	$[\mathbf{dB\mu V}] = [\mathbf{dBm}] + 107$	50Ω system impedance
dBμV	$[\mathbf{dB\mu V}] = [\mathbf{dB\mu A}] + 20\log_{10}(Z)$	Z = system impedance
dBμV	$[\mathbf{dB\mu V}] = [\mathbf{dB\mu A}] + 34$	50Ω system impedance
A	$[\mathbf{A}] = 10^{\left(\frac{[\mathbf{dBA}]}{20}\right)}$	
A	$[\mathbf{A}] = 10^{\left(\frac{([\mathbf{dB\mu A}]-120)}{20}\right)}$	
dBA	$[\mathbf{dBA}] = 20\log_{10}(\mathbf{A})$	
dBA	$[\mathbf{dBA}] = [\mathbf{dB\mu A}] - 120$	
dBμA	$[\mathbf{dB\mu A}] = 20\log_{10}(\mathbf{A}) + 120$	
dBμA	$[\mathbf{dB\mu A}] = [\mathbf{dBm}] - 10\log_{10}(Z) + 90$	Z = system impedance
dBμA	$[\mathbf{dB\mu A}] = [\mathbf{dBm}] + 73$	50Ω system impedance
dBμA	$[\mathbf{dB\mu A}] = [\mathbf{dB\mu V}] - 20\log_{10}(Z)$	Z = system impedance
dBμA	$[\mathbf{dB\mu A}] = [\mathbf{dB\mu V}] - 34$	50Ω system impedance
dBm	$[\mathbf{dBm}] = [\mathbf{dB\mu V}] - 10\log_{10}(Z) - 90$	Z = system impedance
dBm	$[\mathbf{dBm}] = [\mathbf{dB\mu V}] - 107$	50Ω system impedance
dBm	$[\mathbf{dBm}] = [\mathbf{dB\mu A}] + 10\log_{10}(Z) - 90$	Z = system impedance
dBm	$[\mathbf{dBm}] = [\mathbf{dB\mu A}] - 73$	50Ω system impedance
W	$[\mathbf{W}] = 10^{\left(\frac{[\mathbf{dBW}]}{10}\right)}$	
mW	$[\mathbf{mW}] = 10^{\left(\frac{([\mathbf{dBm}])}{10}\right)}$	

Table 3.4 Conversion between field strengths [V/m], [A/m], [dBμV/m], [dBμA/m], [dBpT], and [μT] for free-space (far-field) where $Z_0 = 377\ \Omega$ [1]

[dB] Unit Conversion - Field Strength (free-space, 377Ω)		
To	Calculation	Remark
V/m	$[\mathbf{V/m}] = 10^{\left(\frac{([\mathbf{dB\mu V/m}]-120)}{20}\right)}$	
V/m	$[\mathbf{V/m}] = \sqrt{[\mathbf{W/m^2}] \cdot 377}$	Free space, Z = 377Ω
A/m	$[\mathbf{A/m}] = \frac{[\mathbf{\mu T}]}{1{,}25}$	Free space, Z = 377Ω
dBμV/m	$[\mathbf{dB\mu V/m}] = 20\log_{10}([\mathbf{V/m}]) + 120$	
dBμV/m	$[\mathbf{dB\mu V/m}] = [\mathbf{dBm/m^2}] + 10\log_{10}(Z) + 90$	Z = system impedance
dBμV/m	$[\mathbf{dB\mu V/m}] = [\mathbf{dBm/m^2}] + 115{,}8$	Free space, Z = 377Ω
dBμV/m	$[\mathbf{dB\mu V/m}] = [\mathbf{dB\mu A/m}] + 20\log_{10}(Z)$	Z = system impedance
dBμV/m	$[\mathbf{dB\mu V/m}] = [\mathbf{dB\mu A/m}] + 51{,}5$	Free space, Z = 377Ω
dBμA/m	$[\mathbf{dB\mu A/m}] = [\mathbf{dB\mu V/m}] - 20\log_{10}(Z)$	Z = system impedance
dBμA/m	$[\mathbf{dB\mu A/m}] = [\mathbf{dB\mu V/m}] - 51{,}5$	Free space, Z = 377Ω
dBμA/m	$[\mathbf{dB\mu A/m}] = [\mathbf{dBpT}] - 2$	Free space, Z = 377Ω
dBmW/m²	$[\mathbf{dBm/m^2}] = [\mathbf{dB\mu V/m}] - 10\log_{10}(Z) - 90$	Z = system impedance
dBmW/m²	$[\mathbf{dBm/m^2}] = [\mathbf{dB\mu V/m}] - 115{,}8$	Free space, Z = 377Ω
dBpT	$[\mathbf{dBpT}] = [\mathbf{dB\mu A/m}] + 2$	Free space, Z = 377Ω
μT	$[\mathbf{\mu T}] = [\mathbf{A/m}] \cdot 1{,}25$	Free space, Z = 377Ω

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Chapter 4
Frequency and Wavelength

If you want to find the secrets of the universe, think in terms of energy, frequency and vibration.

—Nikola Tesla

4.1 EMC and Frequencies

In general, *conducted EMC emission and immunity tests* take place at lower frequencies than *radiated EMC emission and immunity tests*.

- **Conducted RF Emissions.** It is assumed that connected cables to electric equipment could act as antennas (in case of conducted emissions from the equipment induced into these cables). Therefore, conducted RF emission limits exist to avoid radiation of connected cables (power, communication). Usually, conducted RF emissions are specified for frequencies $f = 150\,\text{kHz}$ to $30\,\text{MHz}$ [1].
- **Radiated RF Emissions.** Radiated RF emissions tend to occur at higher frequencies (in the megahertz range, depending on the equipment's maximum dimensions [m]). Therefore, regulators usually specify the maximum RF radiation limits for frequencies from $f = 30\,\text{MHz}$ to $f = 6\,\text{GHz}$ [1].
- **Conducted RF Immunity.** As stated above, connected cables are thought to act as antennas. Therefore, conducted RF immunity tests exist to evaluate the functional immunity of electrical and electronic equipment when subjected to conducted disturbances induced to connected cables by RF fields. Conducted RF immunity tests are usually performed with frequencies from $f = 150\,\text{kHz}$ to $80\,\text{MHz}$ [4].
- **Radiated RF Emissions.** Radiated RF immunity tests are intended to demonstrate the immunity of electrical and electronic equipment when subjected to wireless devices like mobile phones and other radiated interference. Radiated RF immunity tests are usually performed with frequencies from $f = 80\,\text{MHz}$ to $6\,\text{GHz}$ [3].
- **Radiated Magnetic Field Immunity.** Magnetic field immunity tests are usually performed at mains power frequency: 50 and 60 Hz [5].

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_4

Table 4.1 ITU frequency bands and their corresponding wavelengths in free-space [11]

Band number	Symbols	Frequency range	Corresponding wave length	Example Uses
1	Extremely low frequency (ELF)	3–30 Hz	100'000–10'000 km	Submarine communication
2	Super low frequenc (SLF)	30–300 Hz	10'000–1'000 km	Submarine communication
3	Ultra low frequency (ULF)	300–3'000 Hz	1'000–100 km	Communication within mines, submarine
4	Very low frequency (VLF)	3–30 kHz	100–10 km	Navigation, time signals, submarine
5	Low frequency (LF)	30–300 kHz	10–1 km	Navigation, time signals, AM, amateur radio
6	Medium frequency (MF)	300–3'000 kHz	1'000–100 m	AM, amateur radio, avalanche beacons
7	High frequency (HF)	3–30 MHz	100–10 m	Shortwave broadcast, amateur radio
8	Very high frequency (VHF)	30–300 MHz	10–1m	FM, TV, aircraft, amateur radio, weather radio
9	Ultra high frequency (UHF)	300–3'000 MHz	1–0.1 m	WiFi, mobile phones, Bluetooth, GPS, TV
10	Super high frequency (SHF)	3–30 GHz	100–10 mm	WiFi, mobile phones, astronomy, satellites
11	Extremely high frequency (EHF)	30–300 GHz	10–1 mm	Astronomy, remote sensing, weapons
12	Tremendously high frequency (THF)	300–3'000 GHz	1–0.1 mm	THz time-domain spectroscopy, tomography

- **Common-Mode Low-Frequency Disturbance**. In some areas, there are also conducted EMC immunity tests specified from $f = 0\,\text{Hz}$ to $f = 150\,\text{kHz}$ [2]. This is intended to demonstrate the immunity of electrical and electronic equipment when subjected to conducted common-mode disturbances such as those originating from power line currents, frequency converters, and return leakage currents in the earthing/grounding system.

The radio spectrum managed by the *International Telecommunication Union (ITU)* goes up to 3000 GHz[1] and is divided into 12 *ITU frequency bands*. Table 4.1) shows the ITU frequency bands with their corresponding wavelength and potential applications.

4.2 Wavelength vs. Frequency

4.2.1 Wavelength in Any Media

The *frequency* f [Hz] of a sinusoidal electromagnetic wave (Fig. 4.1) and its *wavelength* λ [m] have the following relationship [6]:

$$\lambda = \frac{v}{f} = \frac{2\pi}{\beta} = \frac{1}{f\sqrt{\frac{(\epsilon'\mu'-\epsilon''\mu'')}{2}\cdot\left(\sqrt{1+\left(\frac{\epsilon'\mu''+\epsilon''\mu'}{\epsilon'\mu'-\epsilon''\mu''}\right)^2}+1\right)}} \tag{4.1}$$

[1] Radio waves are electromagnetic waves of frequencies arbitrarily lower than 3000 GHz (3 THz), propagated in space without artificial guide. 3 THz is already in the infrared frequency band (300 GHz–430 THz); visible light starts at 430 THz.

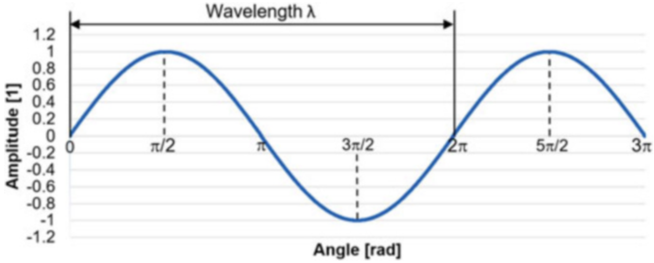

Fig. 4.1 Wavelength λ of a sine wave signal

where:

v = propagation velocity of the signal in [m/sec]
f = frequency of the sinusoidal signal in [Hz]
β = propagation constant in [1/m]; see Eq. 7.53
ϵ' = real part of the complex permittivity $\underline{\epsilon} = \epsilon' - j\epsilon''$ of the medium through which the wave is traveling in [F/m]
ϵ'' = imaginary part of the complex permittivity $\underline{\epsilon} = \epsilon' - j\epsilon''$ of the medium through which the wave is traveling in [F/m]
μ' = real part of the complex permeability $\underline{\mu} = \mu' - j\mu''$ of the medium through which the wave is traveling in [H/m]
μ'' = imaginary part of the complex permeability $\underline{\mu} = \mu' - j\mu''$ of the medium through which the wave is traveling in [H/m]

4.2.2 Wavelength in Insulating Media

In case of an insulator ($\mu_r' = 1$) and negligible dielectric and magnetic losses ($\epsilon'' = 0$, $\mu'' = 0$), the wavelength of a sinusoidal electromagnetic wave with frequency f [Hz] can be written as:

$$\lambda = \frac{v}{f} = \frac{1}{f\sqrt{\mu_0\epsilon_0\epsilon_r'}} = \frac{c}{f\sqrt{\epsilon_r'}} \tag{4.2}$$

where:

v = propagation velocity of the signal in [m/sec]
f = frequency of the sinusoidal signal in [Hz]
$c = 1/(\sqrt{\mu_0\epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity
ϵ_r' = relative permittivity, dielectric constant of the insulator

4.2.3 Wavelength in Vacuum

For vacuum (and approximately air), the calculation of the wavelength of a sinusoidal electromagnetic wave reduces to [10]:

$$\lambda = \frac{c}{f} = \frac{1}{f\sqrt{\mu_0 \epsilon_0}} \tag{4.3}$$

where:

$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
f = frequency of the sinusoidal signal in [Hz]

4.2.4 Wavelength in Good Conducting Media

In case the electromagnetic sinusoidal wave travels through a good conductor (through and not along(!), e.g., through a shield) with negligible magnetic losses ($\mu'' = 0$), the wavelength can be calculated as [10]:

$$\lambda = \sqrt{\frac{4\pi}{f\mu'\sigma}} \tag{4.4}$$

where:

f = frequency of the sinusoidal signal in [Hz]
$\mu' = \mu'_r \mu_0$ = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]
σ = specific conductance of the medium where the wave is propagating through in [S/m]

4.3 Wavelength of Signals Along Wires, Cables, and PCB Traces

It is important to understand that the signal propagation velocity v [m/sec] depends on the transport medium through which the electromagnetic field is traveling. Therefore, the same signal with the same frequency f [Hz] has a different wavelength λ [m] in a blank wire (surrounded by air) than in a cable or PCB trace (surrounded by one or multiple insulation materials). The wavelength λ of signals traveling along wires, cables, and PCB traces—where the dielectric and magnetic losses can be neglected ($\epsilon'' = 0$, $\mu'' = 0$) and the materials around the conductors are assumed to be nonmagnetic ($\mu'_r = 1$)—is given as [9]:

$$\lambda = \frac{v}{f} = \frac{c}{f\sqrt{\epsilon_{reff}}} = \frac{c}{f} \cdot \text{VF} \tag{4.5}$$

where:

v = propagation velocity of the signal in [m/sec]
f = frequency of the sinusoidal signal in [Hz]
$c = 2.998 \cdot 10^8$ m/sec = speed of light
ϵ_{reff} = effective relative permittivity (dielectric constant) of the material(s) through which the electromagnetic field is propagating
VF = velocity factor

4.3.1 Wavelength of Signals Along Blank Wires

The *wavelength* λ [m] of a signal with frequency f [Hz] which travels along a blank wire (or antenna surrounded by air) depends only on the speed of light c [m/sec] and the signal frequency f [Hz] ($v = c$, because $\epsilon_r' = 1$ and $\mu_r' = 1$ and therefore VF = 1) [9]:

$$\lambda_{blankwire} = \frac{c}{f} \tag{4.6}$$

where:

$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
f = frequency of the sinusoidal signal in [Hz]

4.3.2 Wavelength of Signals Along Cables and PCB Traces

The wavelength λ of a signal with frequency f which travels along a wire, *cable*, or a *printed circuit board* (PCB) trace is [9]:

$$\lambda_{cable/PCBtrace} = \frac{c}{f \cdot \sqrt{\epsilon_{reff}}} \tag{4.7}$$

where:

$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
f = frequency of the sinusoidal signal in [Hz]
ϵ_{reff} = the effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating

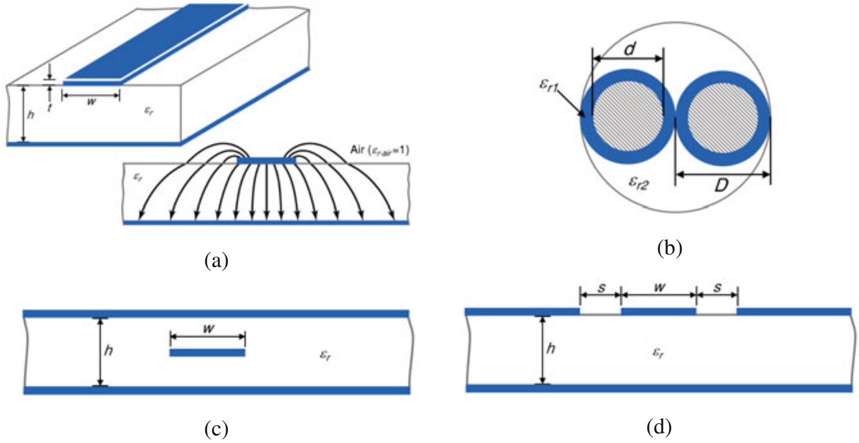

Fig. 4.2 Transmission line examples. (**a**) PCB trace: microstrip line. (**b**) Twisted pair cable. (**c**) PCB trace: stripline. (**d**) PCB trace: coplanar waveguide with reference plane

The effective dielectric constant ϵ_{reff} is defined as the uniform equivalent dielectric constant for a transmission line, even in the presence of different dielectrics (e.g., FR-4 and air for a microstrip line; see Fig. 4.2a). The relative permeability μ_r' is assumed to be equal 1.0 for cables and PCBs, because the insulation materials are nonmagnetic. Thus, the *velocity factor* (VF) depends primarily on the effective relative permittivity ϵ_{reff} through which the electromagnetic wave is propagating.

The calculation of the effective dielectric constant ϵ_{reff} depends on the insulation material and the geometry of the transmission line (e.g., ribbon cable, microstrip, coplanar waveguide, etc.), because the amount of electric field lines in the different media depends on the geometry of the transmission line. Figure 4.2 shows some common transmission lines and Table 4.2 the corresponding ϵ_{reff}

The velocity factor (VF) of a transmission medium is the ratio of the velocity v [m/sec] at which a wavefront of an electromagnetic signal passes through the medium, compared to the speed of light in vacuum c =2.998 $\cdot 10^8$ m/sec:

$$\mathrm{VF} = \frac{v}{c} \tag{4.8}$$

Thus, the smaller the velocity factor (VF), the smaller the wavelength λ [m].

During EMC emission measurement and troubleshooting, it is often necessary to determine the wavelength λ [m] of a certain unintended disturbance with frequency f [Hz] because once you know the wavelength of the disturbance, you can look for potential antennas of the disturbance (e.g., looking for cables with length $l = \lambda/4$ or $l = \lambda/2$ of the disturbance). Table 4.3 presents the velocity factors for different ϵ_{reff} and the resulting wavelength λ [m] for a given frequency f [Hz].

Table 4.2 Approximate velocity factor (VF) for different transmission lines and insulation materials. Calculation of ϵ_{reff} according to: [7, 8, 12]

Material	ε_r	Transmission line	ε_{reff}	VF
FR-4 (fiberglass epoxy of PCBs)	4.5	Microstrip line (w=0.3mm, h=0.5mm)	3.1	0.57
		Stripline (within FR-4 epoxy)	4.5	0.47
		Coplanar waveguide (w=0.3mm, h=0.5mm, s=0.5mm)	3.0	0.58
Polyethylene (PE)	2.26	Twisted pair cable	2.26	0.67
		Ribbon cable	1.3	0.88
Polyvinyl chloride (PVC)	3.0	Twisted pair cable	3.0	0.58
		Ribbon cable	1.5	0.82
Polytetrafluorethylene, Teflon (PTFE)	2.1	Twisted pair cable	2.1	0.69
		Ribbon cable	1.3	0.88
		Coaxial cable	2.1	0.69

Table 4.3 Wavelength λ [m] for given frequencies f [Hz] and dielectric constants ϵ_r'

Frequency [Hz]	λ [m] for free-space ε_r=1	λ [m] for ε_r=1.5, VF=0.82	λ [m] for ε_r=3.0, VF=0.58	λ [m] for ε_r=4.5, VF=0.47
1 Hz	300'000'000	246'000'000	174'000'000	141'000'000
10 Hz	30'000'000	24'600'000	17'400'000	14'100'000
100 Hz	3'000'000	2'460'000	1'740'000	1'410'000
1 kHz	300'000	246'000	174'000	141'000
10 kHz	30'000	24'600	17'400	14'100
100 kHz	3'000	2'460	1'740	1'410
1 MHz	300	246	174	141
10 MHz	30	25	17	14
100 MHz	3.0	2.5	1.7	1.4
1 GHZ	0.30	0.25	0.17	0.14
10 GHz	0.030	0.025	0.017	0.014
100 GHz	0.0030	0.0025	0.0017	0.0014
1 THz	0.00030	0.00025	0.00017	0.00014

4.3.3 Summary

- **Wavelength.** The wavelength λ [m] of a sinusoidal signal with frequency f [Hz] depends on the media through which the electromagnetic wave is propagating because the velocity v [m/sec] changes with the dielectric and magnetic properties $\underline{\epsilon}$ [F/m], $\underline{\mu}$ [H/m].
- **Wavelength of signals along conductors.**

$$\lambda = v/f = c/(f\sqrt{\epsilon_{reff}}) \tag{4.9}$$

where:

v = propagation velocity of the signal in [m/sec]
f = frequency of the sinusoidal signal in [Hz]
ϵ_{reff} = the effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating

- **Wavelength of electromagnetic waves in free-space.**

$$\lambda = c/f \tag{4.10}$$

where:

$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
f = frequency of the sinusoidal electromagnetic wave in [Hz]

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Chapter 5
Time-Domain and Frequency-Domain

Fourier's theorem has all the simplicity and yet more power than other familiar explanations in science. Stated simply, any complex pattern, whether in time or space, can be described as a series of overlapping sine waves of multiple frequencies and various amplitudes.

—Bruce Hood

5.1 Fourier Analysis

Figure 5.1 presents measurement instruments for measuring electrical signals in the time-domain (oscilloscope) and the frequency-domain (spectrum analyzer). *Fourier analysis* refers to the mathematical principle that every signal can be represented by the sum of simple trigonometric functions (sine, cosine, etc.). The Fourier analysis enables a transformation of a signal in the time-domain $x(t)$ to a signal in the frequency-domain $X(\omega)$, where $\omega = 2\pi f$. In other words, a Fourier analysis is a mathematical operation for calculating the frequency-domain representation (frequency spectrum) of a signal in the time-domain. Two common Fourier analysis notations are:

$$x(t) \circ\!\!-\!\!\bullet X(\omega)$$

$$\mathscr{F}\{x(t)\} = X(\omega)$$

where:

$x(t)$ = a signal in the time-domain
$X(\omega)$ = the Fourier transform of $x(t)$ (frequency-domain)

Figure 5.2 shows a representation of a square wave signal (1 V amplitude) with the sum of only four harmonic sine waves (first, third, fifth, and seventh) and a direct current (DC) component of 0.5 V. For representing an ideal square wave, an

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_5

(a)

(b)

Fig. 5.1 Time-domain and frequency-domain measurement equipment. However, some oscilloscopes do also have built-in Fourier analysis functionality (FFT). (**a**)Time domain: high-end oscilloscope R&S®RTO64 by Rohde & Schwarz. (**b**) Frequency domain: signal and spectrum analyzer R&S®FSVA3030 by Rohde & Schwarz

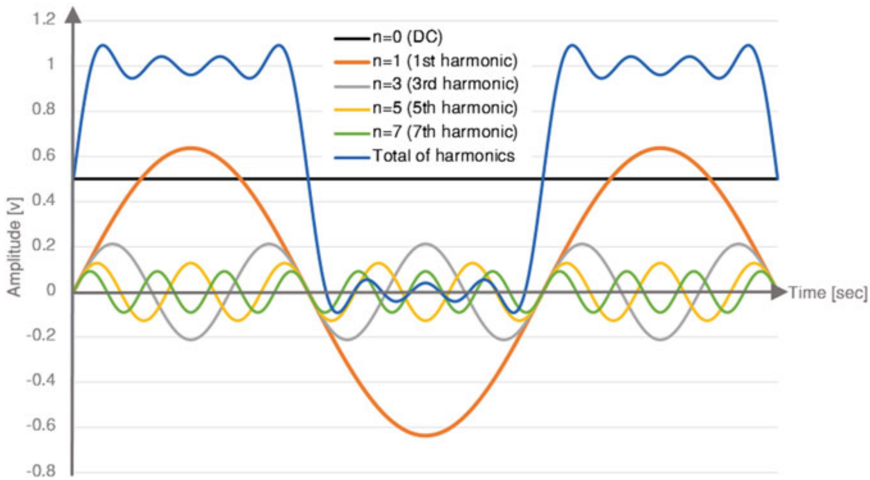


Fig. 5.2 Fourier analysis of a square wave; up to the seventh harmonic

indefinite number of sine waves would be necessary (because the rise- and fall-time of an ideal square wave are 0 s).

There are different types of integral transforms. These transforms have in common that they define the necessary math for converting a signal from time- to the frequency-domain and vice versa. Every integral transform has its field of application:

- **Fourier Series.** Continuous and periodic signals.
- **Fourier Transform.** Continuous, nonperiodic signals.
- **Discrete Fourier Transform (DFT).** Discrete and periodic signals.

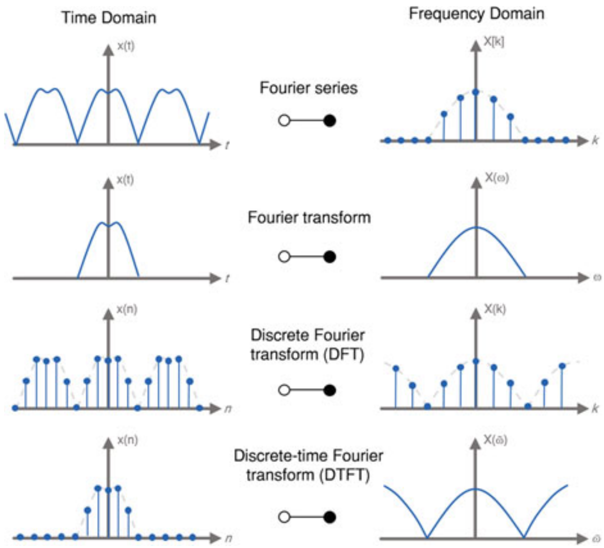


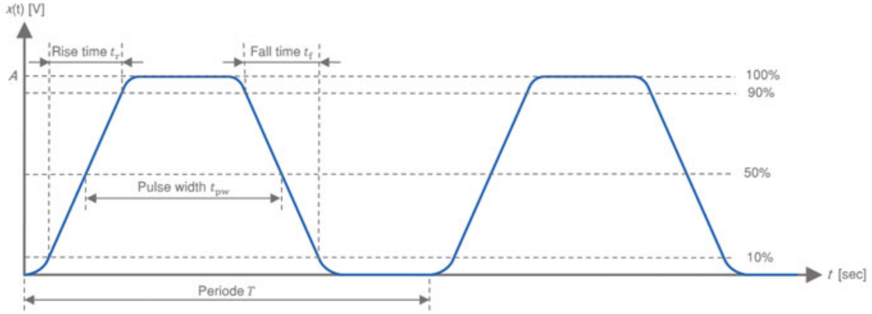
Fig. 5.3 Time-domain to frequency-domain transforms

- **Fast Fourier Transform (FFT).** The FFT is a DFT. The FFT is an implementation of the DFT for fast and efficient computation.
- **Discrete-time Fourier Transform (DTFT).** Discrete, nonperiodic signals.
- **Laplace Transform.** Control systems and filter design.
- **Z-Transform.** Time-discrete control systems and filter design.

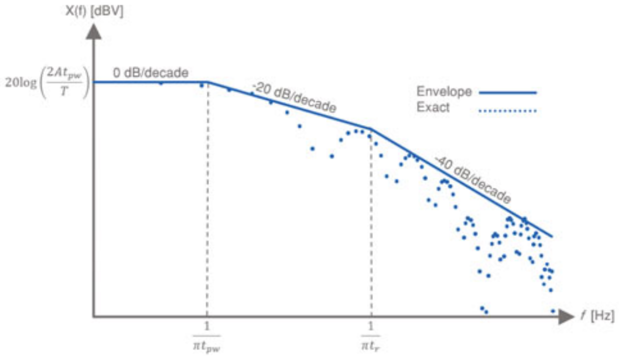
Figure 5.3 shows some common integral transforms used in the field of EMC. Further details about the different integral transforms can be found in the Appendix F.

5.2 Frequency Spectra of Digital Signals

Digital signals like clocks and data interfaces are the primary cause for EMC emissions of electronic circuits and electrical equipment and systems. Digital signals in the time-domain can be represented by trapezoid-shaped pulses with a period time T [sec], a pulse width t_{pw} [sec], a rise-time t_r [sec], and a fall-time t_f [sec]. Figure 5.4 shows an extract of a digital waveform and the corresponding frequency spectrum (Fourier analysis). It can be seen that the pulse width t_{pw} [sec] defines the frequency at which the amplitude spectrum starts to drop with -20 dB/decade, whereas the minimum fall- and/or rise-time defines the frequency at which the spectrum starts to drop with -40 dB/decade. Figure 5.5a compares the frequency spectrum of clock signals with different duty-cycles [%] (pulse



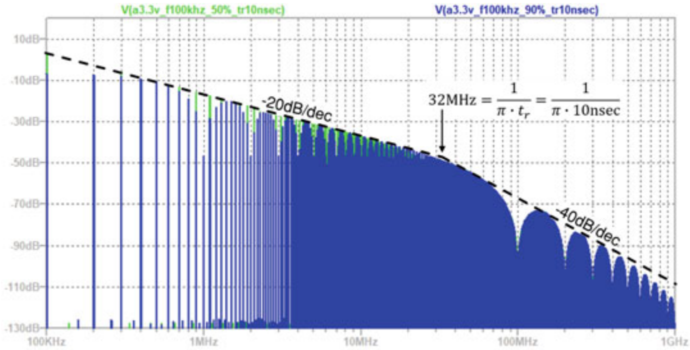
(a)



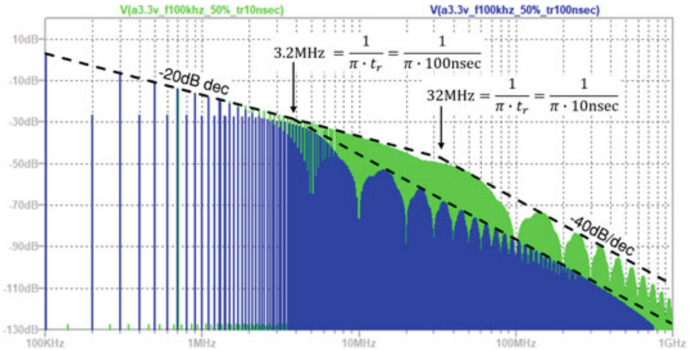
(b)

Fig. 5.4 Digital signal in the time- and frequency-domain. (**a**) Digital waveform with period time T [sec], pulse width t_{pw} [sec], rise-time t_r [sec] and fall-time t_f [sec]. (**b**) Amplitude frequency spectrum of a digital signal and its envelope curve [3]

widths). It is remarkable that a clock signal with a 90 % duty-cycle has a lower amplitude of the first harmonic than a clock signal with a 50 % duty-cycle (of the same frequency [Hz] and with the same rise-/fall-time [sec]). However, as a 90 % clock signal has more power, this power adds to the DC component (0 Hz). For EMC design engineers, Fig. 5.5b is even more important than Fig. 5.5a because it shows how high-frequency harmonics of a clock signal can be reduced effectively. In this example, an increase of the rise- and fall-time [sec] by factor 10 reduces the amplitude of the high-frequency harmonics ($f > 32$ MHz) also by factor 10 (20 dB).



(a)



(b)

Fig. 5.5 Clock signal parameter analysis in the frequency-domain [1]. (**a**) Frequency spectrum of 100 kHz clock signals with rise-/fall-times of 10 nsec and duty-cycles of 50% vs. 90%. (**b**) Frequency spectrum of 100 kHz clock signals with 50% duty-cycle and rise-/fall-times of 10 nsec vs. 100 nsec

5.3 Bandwidth of Digital Signals

Bandwidth [Hz] of a digital signal means what is the highest significant sine wave frequency component in the digital signal? Significant in this case means that the power in [W] in the frequency component is bigger than 50 % of the power in an ideal square wave's signal with the same amplitude A in [V] and duty-cycle $D = (t_{pw}/T)$ in [%]. A drop in 50 % of the power [W] is the same as a drop of 70 % in amplitude [V] or a drop of 3 dB.

The rule of thumb for calculating the bandwidth [Hz]—or the highest significant sine wave frequency—of a trapezoid digital signal is [2]:

$$B = \frac{0.35}{t_{10-90\%}} \tag{5.1}$$

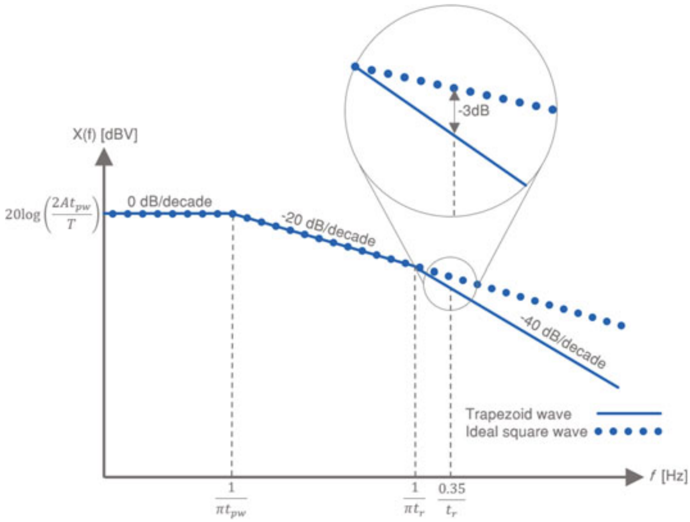


Fig. 5.6 The 3dB-bandwidth of a clock signal with a pulse width of t_{pw} [sec], rise- and fall-times of $t_r = t_f$ [sec]

where:

B = bandwidth = highest significant sine wave frequency (harmonic) in a digital signal [Hz]

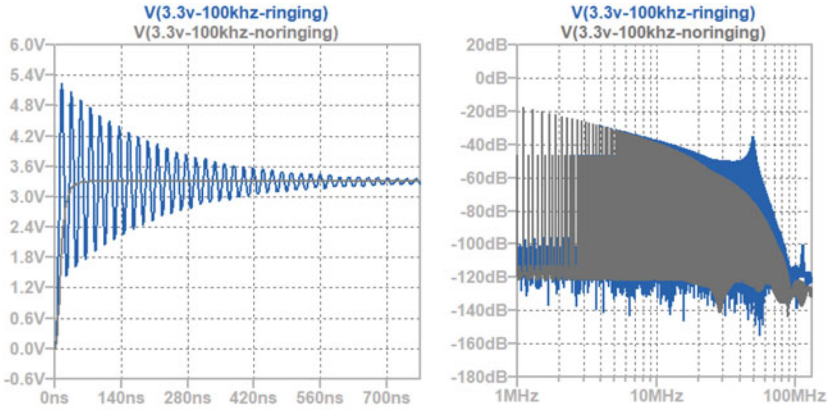
$t_{10-90\%}$ = rise- and/or fall-time (whichever is smaller) from 10 % to 90 % of the slope of a digital signal in [sec]

Figure 5.6 shows the frequency spectrum envelop curves of an ideal square wave $t_r = t_f = 0$ sec and a trapezoid with $t_r > 0$ sec and $t_f > 0$ sec. The 3 dB-bandwidth of the trapezoid waveform can be found at $f_{3dB} = 0.35/t_{10-90\%}$.

Note: it is assumed that there is no ringing in the waveform and $t_r = t_f$. In case of ringing, the frequency spectrum envelope for $f > 1/(\pi t_r)$ would not drop off with −40 dB/decade. Instead, an increase in the frequency spectrum is assumed at the ringing frequency (see Fig. 5.7).

5.4 Ringing and Frequency Spectrum

Impedance mismatch along a transmission line could cause *ringing*. One could argue that ringing has nothing to do with EMC and that ringing is a signal integrity topic. However, this is not entirely correct. Due to ringing, higher frequency components (harmonics) are introduced, and it may be possible that the conductor—where the signal is propagating along—acts as an efficient antenna at these higher frequencies.



(a) (b)

Fig. 5.7 Simulation data of a signal with ringing (blue) vs. without ringing (gray). Both signals have an amplitude of 3.3 V, a fundamental frequency of 100 kHz, and a rise-/fall-time of 10 nsec [1]. (**a**) Time-domain. (**b**) Frequency-domain

In the example of Fig. 5.7, the harmonics around 50 MHz show as much as 100 times higher voltage amplitudes (40 dB) in the signal with ringing than compared to the signal without ringing. This example illustrates that impedance matching does also matter for EMC and that EMC design engineers must know which signal connections must be considered as transmission lines and which not.

5.5 Summary

- **Time- vs. frequency-domain.** Electrical signals have a time-domain $x(t)$ and a frequency-domain $X(\omega)$ representation.
- **Measurement methods.** Time-domain representations are typically measured with an oscilloscope and frequency-domain representations with a spectrum analyzer.
- **Bandwidth of digital signals.**

$$B \approx 0.35/t_{10-90\%} \tag{5.2}$$

where:

B = highest significant sine wave frequency (harmonic) in a digital signal [Hz]
$t_{10-90\%}$ = rise- and/or fall-time (whichever is smaller) from 10 % to 90 % of the digital signal in [sec]

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Chapter 6
RF Parameters

The term impedance matching can be a very useful metaphor for connoting important aspects of social interactions. For example, the smooth and efficient functioning of social networks, whether in a society, a company, a group activity, and especially in relationships such as marriages and friendships, requires good communication in which information is faithfully transmitted between groups and individuals. When information is dissipated or "reflected," such as when one side is not listening, it cannot be faithfully or efficiently processed, inevitably leading to misinterpretation, a process analogous.

—Geoffrey West

6.1 Reflection Coefficient $\underline{\Gamma}$

We speak of matched impedances in case the load impedance $\underline{Z}_{load}$ is the complex conjugate of the source impedance $\underline{Z}_{source} = \underline{Z}^*_{load}$ (Fig. 6.1). In radiated emission and immunity EMC testing, it is important to understand the term matching and how to quantify it. All EMC RF measurement test setup receiver and transmitter antennas must be matched to their receiver and transmitter equipment impedance (typically $Z_0 = 50\,\Omega$).

The *reflection coefficient* $\underline{\Gamma}$ is defined as [6]:

$$\underline{\Gamma} = \frac{\underline{V}_{reflected}}{\underline{V}_{forward}} \tag{6.1}$$

$$\underline{\Gamma} = \frac{\underline{Z}_{load} - \underline{Z}_{source}}{\underline{Z}_{load} + \underline{Z}_{source}} \tag{6.2}$$

where:

$\underline{V}_{forward}$ = complex forward voltage wave to the load in [V]
$\underline{V}_{reflected}$ = complex reflected voltage wave by the load in [V]

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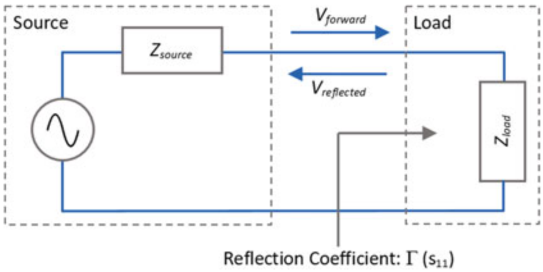


Fig. 6.1 A *one-port* circuit and its reflection coefficient $\underline{\Gamma}$. For a one-port $\underline{\Gamma}$ is equal to the scattering parameter $\underline{s}_{11}$

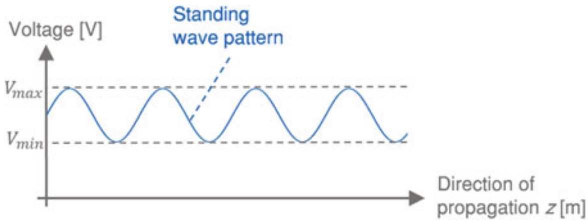


Fig. 6.2 Example of a standing wave pattern along a transmission line (envelope curve of the sinusoidal forward wave $\underline{V}_{forward}$ and its reflected wave $\underline{V}_{reflected}$ of the same frequency)

$\underline{Z}_{source}$ = complex source impedance in [Ω]
$\underline{Z}_{load}$ = complex load impedance in [Ω]

The reflection coefficient Γ is often given in [dB]:

$$\Gamma \text{ [dB]} = 10 \cdot \log_{10}\left(\frac{P_{reflected}}{P_{forward}}\right) \tag{6.3}$$

where:

$P_{reflection}$ = reflected power by the load in [W]
$P_{forward}$ = power sent to the load (at the load terminals) in [W]

6.2 Voltage Standing Wave Ratio (VSWR)

VSWR is dimensionless and means voltage standing wave ratio. The VSWR expresses the ratio of maximum and minimum voltage of a standing wave pattern along a transmission line. It is also a parameter for measuring the degree of impedance matching. Standing waves occur in case of impedance mismatch. A VSWR value of 1 means perfectly matched. A VSWR value of ∞ means complete mismatch (100% of the forward wave is reflected). Figure 6.2 shows an example of a standing wave pattern.

The VSWR can be calculated by using the reflection coefficient from above [6]:

$$\mathrm{VSWR} = \frac{|V|_{max}}{|V|_{min}} = \frac{1 + |\underline{\Gamma}|}{1 - |\underline{\Gamma}|} \tag{6.4}$$

where:

$|V|_{max}$ = is the maximum value of the standing wave pattern in [V]
$|V|_{min}$ = is the minimum value of the standing wave pattern in [V]
$|\underline{\Gamma}|$ = is the magnitude of the reflection coefficient

6.3 Return Loss (RL)

The *return loss* RL [dB] is a measure for the transferred power to a load by a source (Fig. 6.3). A low RL value indicates that not much power is transferred to the load and is reflected instead. Return loss [dB] is always a positive value and it is equal the reflection coefficient Γ [dB] multiplied by -1 [5, 9].

$$\mathrm{RL}\ [\mathrm{dB}] = 10 \cdot \log_{10}\left(\frac{P_{forward}}{P_{reflected}}\right) = -20 \cdot \log_{10}\left(|\underline{\Gamma}|\right) \tag{6.5}$$

$$\mathrm{RL}\ [\mathrm{dB}] = -\Gamma[\mathrm{dB}] \tag{6.6}$$

where:

$P_{forward}$ = power sent to the load (at the load terminals) in [W]
$P_{reflection}$ = reflected power by the load in [W]
$|\underline{\Gamma}|$ = is the magnitude of the reflection coefficient

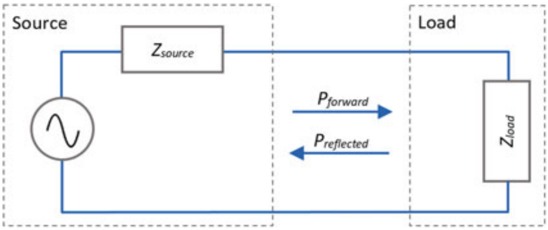


Fig. 6.3 Source with forward power $P_{forward}$ and the reflected power $P_{reflected}$ from a one-port circuit

6.4 Insertion Loss (IL)

The term *insertion loss* IL [dB] is generally used for describing the amount of power loss due to the insertion of one or several of the following components (passive two-port networks):

- Transmission line (cable, PCB trace)
- Connector
- Passive filter

The insertion loss is the ratio in [dB] of the power P_1 and P_2 in Fig. 6.4. P_1 represents the power, which would be transferred to the load in case the source is directly connected to the load. The power P_2 represents the power, which is transferred to the load in case the passive two-port network is inserted between the source and the load [2, 6].

$$\mathrm{IL}\ [\mathrm{dB}] = 10 \cdot \log_{10}\left(\frac{P_1}{P_2}\right) = -10 \cdot \log_{10}\left(\frac{P_2}{P_1}\right) \tag{6.7}$$

where:

P_1 = power, which would be transferred to the load in case the source is directly connected to the load (without passive two-port network) in [W]
P_2 = power, which is transferred to the load in case the passive two-port network is inserted between the source and the load in [W]

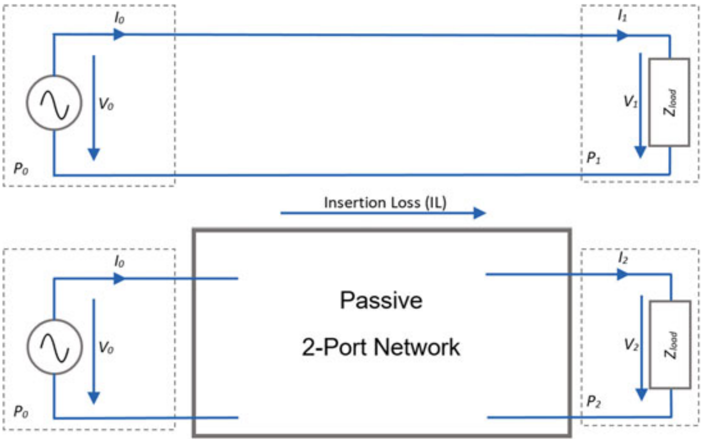


Fig. 6.4 Insertion loss

6.5 Scattering Parameters

Scattering parameters —also called S-parameters—are commonly used in high-frequency or microwave engineering to characterize a *two-port circuit* (see Figs. 6.5 and 6.6). The scattering parameters describe the relation of the power wave parts $\underline{a}_1$, $\underline{b}_1$, $\underline{a}_2$, $\underline{b}_2$ that are transferred and reflected from a two-port input and output. The physical dimension for the incident $\underline{a}$ and reflected $\underline{b}$ power waves is not Watt, but it is $\sqrt{\text{Watt}}$. The generic definitions for the forward power wave $\underline{a}_i$ and the reflected power wave $\underline{b}_i$ of an arbitrary port i are given as [1]:

$$\underline{a}_i = \frac{\underline{V}_i + \underline{Z}_i \underline{I}_i}{2\sqrt{|\operatorname{Re}(\underline{Z}_i)|}} \tag{6.8}$$

$$\underline{b}_i = \frac{\underline{V}_i - \underline{Z}_i^* \underline{I}_i}{2\sqrt{|\operatorname{Re}(\underline{Z}_i)|}} \tag{6.9}$$

where:

$\underline{V}_i$ = complex voltage at the input of the ith port of a junction in [V]
$\underline{I}_i$ = current flowing into the ith port of a junction in [A]
$\underline{Z}_i$ = complex impedance of the ith port in [Ω]
$\underline{Z}_i^*$ = complex conjugate of $\underline{Z}_i$
The positive real value is chosen for the square root in the denominators

The forward traveling power toward port 1 P_{1fwd} and the reflected power of port 1 P_{1ref} can be written as:

$$P_{1fwd} = \frac{1}{2} \cdot |\underline{a}_1|^2 = \frac{|\hat{V}_{1fwd}|^2}{2\operatorname{Re}(\underline{Z}_1)} \tag{6.10}$$

$$P_{1ref} = \frac{1}{2} \cdot |\underline{b}_1|^2 = \frac{|\hat{V}_{1ref}|^2}{2\operatorname{Re}(\underline{Z}_1)} \tag{6.11}$$

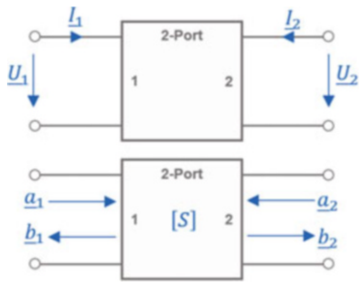


Fig. 6.5 Scattering parameters of a two-port network

Fig. 6.6 S-parameters are measured with *vector network analyzers*. The picture shows the R&S®ZNL6 by Rohde & Schwarz

where:

$\hat{V}_{1fwd}$ = peak value of the forward voltage to port 1 in [V]
$\hat{V}_{1ref}$ = peak value of the reflected voltage from port 1 in [V]
$\mathrm{Re}(\underline{Z}_1)$ = real part of the complex impedance $\underline{Z}_1$ of port 1 in [Ω]

Generally speaking, the S-parameter $\underline{s}_{ij}$ is determined by driving port j with an incident wave of voltage V_j^+ and measuring the outgoing voltage wave V_i^- at port i. Considering Fig. 6.5, the four scattering parameters can be computed as follows:

$$\underline{s}_{11} = \left.\frac{\underline{b}_1}{\underline{a}_1}\right|_{\underline{a}_2=0} = \text{input reflection coefficient} \tag{6.12}$$

$$\underline{s}_{21} = \left.\frac{\underline{b}_2}{\underline{a}_1}\right|_{\underline{a}_2=0} = \text{forward gain, forward transmission} \tag{6.13}$$

$$\underline{s}_{12} = \left.\frac{\underline{b}_1}{\underline{a}_2}\right|_{\underline{a}_1=0} = \text{backward gain, reverse transmission} \tag{6.14}$$

$$\underline{s}_{22} = \left.\frac{\underline{b}_2}{\underline{a}_2}\right|_{\underline{a}_1=0} = \text{output reflection coefficient} \tag{6.15}$$

where:

$\underline{a}_1$ = incoming power wave at the input port of a two-port in $[\sqrt{W}]$
$\underline{b}_1$ = outgoing power wave at the input port of a two-port in $[\sqrt{W}]$
$\underline{a}_2$ = incoming power wave at the output port of a two-port in $[\sqrt{W}]$
$\underline{b}_2$ = outgoing power wave at the output port of a two-port in $[\sqrt{W}]$

The formulas 6.12, 6.14, 6.13, 6.15 showed that the S-parameters can be determined by loading the ports with a reference impedance, e.g., $Z_0 = 50\,\Omega$. This means that $\underline{s}_{11} = \underline{\Gamma}$ when the output port 2 is matched with its load. For a one-port $\underline{s}_{11}$ is always equal the reflection coefficient $\underline{\Gamma}$. The input reflection coefficient $\underline{\Gamma}_1$ of a two-port with a certain load (given by the reflection factor $\underline{\Gamma}_L$) is given by:

$$\underline{\Gamma}_1 = \underline{s}_{11} + \frac{\underline{s}_{12}\underline{s}_{21}\underline{\Gamma}_L}{1 - \underline{s}_{22}\underline{\Gamma}_L} \tag{6.16}$$

6.6 Signal-to-Noise Ratio

The *signal-to-noise-ratio* SNR [dB] is a key figure in analog or digital signal processing applications. The bigger the SNR value, the better. The SNR is defined as the ratio of the desired signal power to the power of the undesired signal (noise, interference) [3]:

$$\text{SNR} = \frac{P_{signal}}{P_{noise}} \tag{6.17}$$

$$\text{SNR [dB]} = 10 \log_{10} \left(\frac{P_{signal}}{P_{noise}} \right) \tag{6.18}$$

where:

P_{signal} = power of the desired signal in [W]
P_{noise} = power of the undesired signal (noise, interference) in [W]

6.7 Noise Factor and Noise Figure

Noise factor F and *noise figure* NF [dB] are measures of degradation of the signal-to-noise ratio (SNR), caused by components in a signal chain. The noise factor of a device (e.g., amplifier) is defined as the ratio of the SNR at the input of the device to the SNR at the output of the device [3]:

$$F = \frac{\text{SNR}_i}{\text{SNR}_o} \tag{6.19}$$

where:

SNR_i = SNR of the signal at the input of the device (linear, not in [dB])
SNR_o = SNR of the signal at the output of the device (linear, not in [dB])

The noise figure NF [dB] is defined as the noise factor F in [dB]:

$$\text{NF} = 10 \log_{10} (F) = 10 \log_{10} \left(\frac{\text{SNR}_i}{\text{SNR}_o} \right) \tag{6.20}$$

where:

F = noise factor of the device (linear, not in [dB])
SNR_i = SNR of the signal at the input of the device (linear, not in [dB])
SNR_o = SNR of the signal at the output of the device (linear, not in [dB])

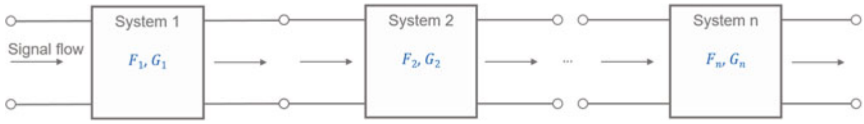


Fig. 6.7 Cascade of devices with gain G_n and noise factor F_n

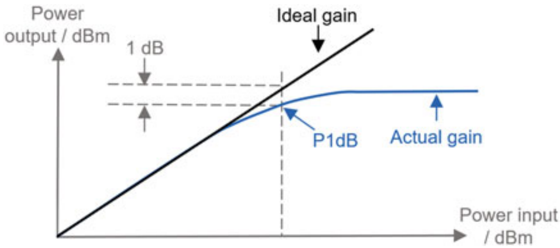


Fig. 6.8 1 dB compression point (P1dB) of an amplifier

If several devices are cascaded (see Fig. 6.7), the total noise factor F can be calculated with Friis formula [3]:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \cdots + \frac{F_n - 1}{G_1 G_2 G_3 \cdots G_{n-1}} \tag{6.21}$$

where:

F_n = noise figure of the n-th device (linear, not in [dB])
G_n = gain of the n-th device (linear, not in [dB])

The first amplifier in a chain usually has the most significant effect on the total noise figure NF. Equation 6.21 shows that the first device should have the lowest noise factor F_1 and a high gain G_1, since the noise figure of any subsequent stage will be divided by the gains of the preceding stages. As a consequence, the noise figure requirements of subsequent stages are more relaxed.

The noise figure NF [dB] of a passive component (cable, attenuator, filter) is always its loss L, and one can set $F = L$ and $G = 1/L$ (linear, not in [dB]) for a passive component [3], where G is the gain.

6.8 1 dB Compression Point

The 1 dB compression point P1dB [dB] is a key specification for amplifiers. Besides many factors, the gain G [dB] of an amplifier is a function of the amplifier's input power P_{in} [dBm]. P1dB is defined as the output power P_{out} [dBm] (or sometimes as the input power P_{in}) of an amplifier at which the output is 1 dB lower than it is supposed to be, if it were ideal. In other words, P1dB is the output power P_{out} [dBm] when the amplifier is at the 1 dB compression point (Fig. 6.8).

Once an amplifier reaches its P1dB, it goes into compression and becomes a nonlinear device, producing distortion, harmonics, and intermodulation products. Therefore, amplifiers should always be operated below the P1dB.

6.9 Spectrum Analyzer Terms

Spectrum analyzers measure the power of a signal's frequency components. Today, spectrum analyzers apply these measurement principles:

- **Superheterodyne.** The analog measurement signal is mixed with a sweep generator frequency signal to the intermediate frequency IF, then band-pass filtered with the resolution bandwidth RBW (IF filter) and finally measured (envelop detector), low-pass filtered (VBW filter), and displayed (see Fig. 6.9a).
- **FFT.** The analog measurement signal is sampled and converted to a digital signal, where the FFT algorithm is used to calculate the frequency spectrum of that measurement signal (see Fig. 6.9a).

Basically, there are three types of spectrum analyzers:

- **Swept-tuned analyzers** apply the principle of superheterodyne receivers. They have long scan times and a wide frequency range.
- **FFT analyzers** apply the principle of the FFT. They have a limited frequency range and short scan times.

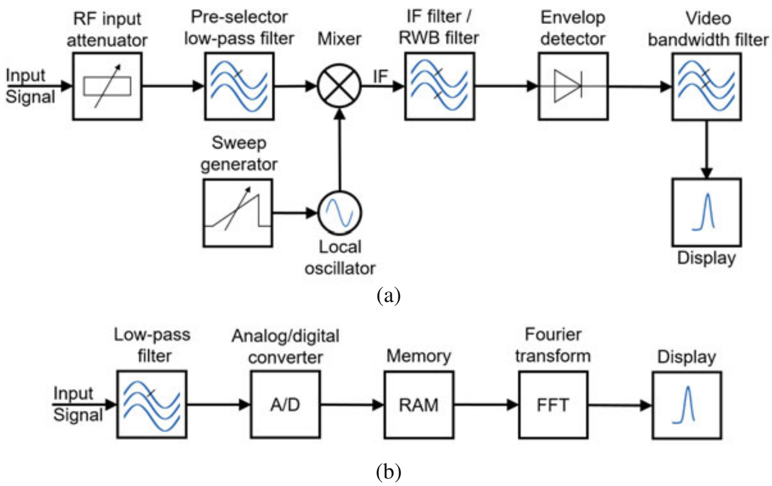


Fig. 6.9 Block diagram of spectrum analyzer principles . (**a**) Superheterodyne principle. (**b**) FFT principle

- **Hybrid superheterodyne-FFT analyzers** combine the principle of superheterodyne measurement and FFT calculation. They have shorter scan times than purely swept-tuned analyzers and a wider bandwidth then FFT analyzers.

Sections 6.9.1–6.9.5 explain the basic terms of spectrum analyzers. Note: *EMI receivers* are spectrum analyzers that fulfill the requirements according to CISPR 16-1-1 [8] (dynamic range, peak-detection, etc.).

6.9.1 Frequency Range

A spectrum analyzer measures the input signal from the *start frequency* to the *stop frequency*. The frequency between the start and the stop frequency is called *center frequency*, and the difference is the frequency *span*.

$$\text{center frequency} = (\text{start frequency} + \text{stop frequency})\,/2 \tag{6.22}$$

$$\text{span} = \text{stop frequency} - \text{start frequency} \tag{6.23}$$

6.9.2 Resolution Bandwidth

For spectrum analyzers that employ the superheterodyne principle, the *resolution bandwidth* (RBW) is the bandwidth of the band-pass IF filter in Fig. 6.9a. The RBW determines how close two signals can be resolved into two separate peaks. The smaller the RBW, the better the resolution in the frequency range and the lower the noise floor (see Fig. 6.10).

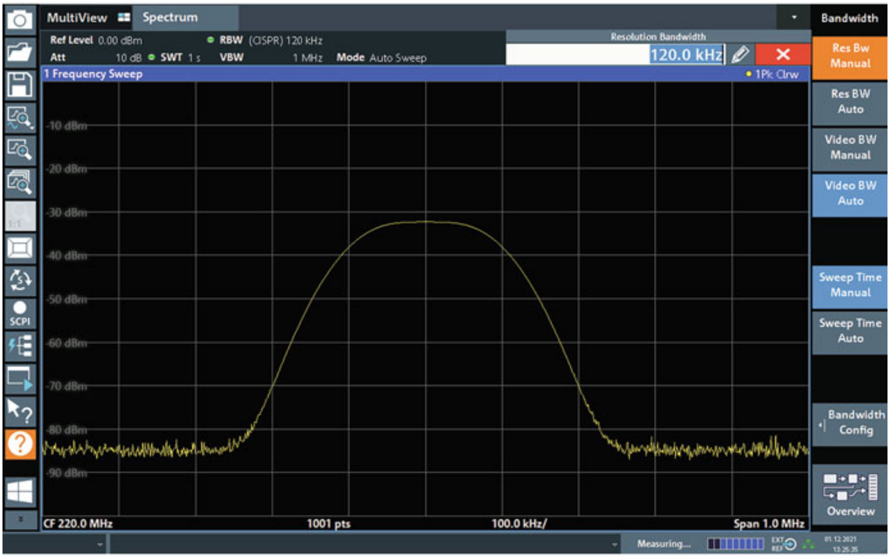
6.9.3 Video Bandwidth

For spectrum analyzers that employ the superheterodyne principle, the *video bandwidth* (VBW) is the bandwidth of the low-pass filter after the envelope detector in Fig. 6.9a. The VBW is used to discriminate between a low-power harmonic signal and the noise floor. The smaller the VBW, the smoother the envelope signal and the lower the noise floor (see Fig. 6.11).

6.9.4 Sweep Time

For spectrum analyzers that employ the superheterodyne principle, the *sweep time* t_{sweep} [sec] is the time required to record the entire frequency range (span) [4]:

(a)



(b)

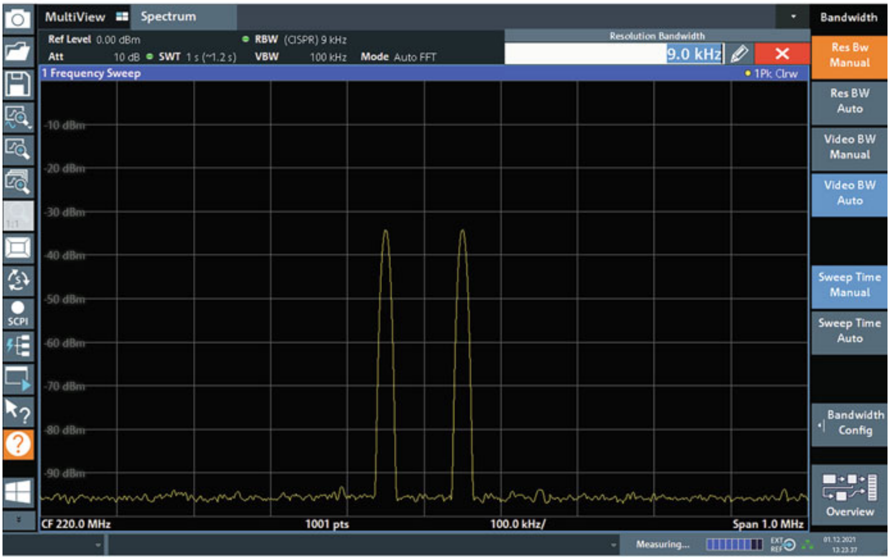


Fig. 6.10 Measurement of the same signal with different resolution bandwidths (RBWs). Measurements performed with R&S®FSV3013 by Rohde & Schwarz. (**a**) RBW = 120 kHz. (**b**) RBW = 9 kHz

(a)

(b)

Fig. 6.11 Measurement of the same signal with different video bandwidths (VBWs). Measurements performed with R&S®FSV3013 by Rohde & Schwarz. (**a**) VBW = 200 kHz. (**b**) VBW = 1 Hz

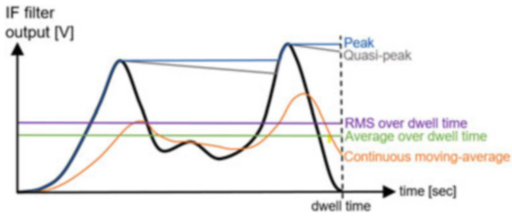


Fig. 6.12 Spectrum analyzer detector types compared (symbolic picture)

$$t_{sweep} = \begin{cases} \frac{k \cdot \text{span}}{\text{RBW}^2} & \text{when VBW} > \text{RBW} \\ \frac{k \cdot \text{span}}{\text{RBW} \cdot \text{VBW}} & \text{when VBW} \leq \text{RBW} \end{cases} \tag{6.24}$$

where:

k = Factor, depending on the type of filter and on the required level of accuracy
span = frequency range to be displayed in [Hz]
RBW = resolution bandwidth in [Hz]
VBW = video bandwidth in [Hz]

6.9.5 Detectors

Different detector types are used in spectrum analyzers (Fig. 6.12):

- **Quasi-peak detector.** The *quasi-peak detector* determines the peaks of the measurement signal with a defined charge and discharge time [4]. Quasi-peak-detection mode is used for many RF emission measurements.
- **Average detector.** The *average detector* determines the linear average of the signal at the output of the IF envelope detector [7]. Many EMC RF emission measurements use average detectors.
- **Peak detector.** The *peak detector* follows the signal at the output of the IF envelope detector and holds the maximum value during the measurement time (also called dwell time) until its discharge is forced [7].
- **RMS detector.** The *RMS detector* determines the RMS value of the signal at the output of the IF envelope detector [7].

6.10 Summary

Table 6.1 summarize the impedance matching parameters and Table 6.2 compares the impedance matching parameters for matched vs. unmatched.

Table 6.1 Impedance matching parameters. Z_0 [Ω] denotes the system impedance. Typically, the system impedance is $Z_0 = 50\,\Omega$ (RF), $Z_0 = 60\,\Omega$ (CAN-bus), $Z_0 = 75\,\Omega$ (television) or $Z_0 = 100\,\Omega$ (Ethernet)

	$f(Z_{load})$	$f(\Gamma)$	f(VSWR)	f(RL[dB])
Z_{load}	–	$\frac{1+\Gamma}{1-\Gamma} \cdot Z_0$	$\mathrm{VSWR}^{\pm 1} \cdot Z_0$	$\left\lvert\frac{10^{-\frac{RL}{20}}+1}{10^{-\frac{RL}{20}}-1}\right\rvert^{\pm 1} \cdot Z_0$
Γ	$\frac{Z_{load}-Z_0}{Z_{load}+Z_0}$	–	$\pm\frac{\mathrm{VSWR}-1}{\mathrm{VSWR}+1}$	$\pm 10^{-\frac{RL}{20}}$
VSWR	$\max\left(\frac{Z_{load}}{Z_0}, \frac{Z_0}{Z_{load}}\right)$	$\frac{1+\lvert\Gamma\rvert}{1-\lvert\Gamma\rvert}$	–	$\left\lvert\frac{10^{-\frac{RL}{20}}+1}{10^{-\frac{RL}{20}}-1}\right\rvert$
RL	$20 \cdot \log_{10}\left(\left\lvert\frac{Z_{load}+Z_0}{Z_{load}-Z_0}\right\rvert\right)$	$-20 \cdot \log_{10}(\lvert\Gamma\rvert)$ $-\Gamma$ [dB]	$20 \cdot \log_{10}\left(\frac{\mathrm{VSWR}+1}{\mathrm{VSWR}-1}\right)$	–

Table 6.2 Qualitative analysis of impedance matching

Degree of Matching	Z_{load} [Ω]	|Γ| [1]	|Γ| [dB]	|VSWR|:1 [1]	RL [dB]	Power Transfer [%]	Power Reflection [%]
Perfectly matched	$Z_{load} = Z_0$	0	-Infinite (∞)	1:1	Infinite (∞)	100	0
Good matching	$Z_{load} = 0.7 \cdot Z_0$ $Z_{load} = 1.4 \cdot Z_0$	0.2	>−14 [dB]	< 1.5:1	> 14 [dB]	> 96	< 4
Poor matching	$Z_{load} = 0.5 \ldots 0.7 \cdot Z_0$ $Z_{load} = 1.4 \ldots 1.9 \cdot Z_0$	0.2...0.3	-14...−10 [dB]	1.5:1...2:1	10...14 [dB]	89...96	4...11
Unmatched	$Z_{load} < 0.5 \cdot Z_0$ $Z_{load} > 1.9 \cdot Z_0$	> 0.3	0...−10 [dB]	> 2:1	0...10 [dB]	< 89	> 11
Total reflection	$Z_{load} = 0$ $Z_{load} = \infty$	1	0	Infinite (∞)	0	0	100

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Chapter 7
Transmission Lines

In the mid-1980s or so, connecting conductors were of no consequence; i.e., the voltage and current at the input to the line are almost identical to the voltage and current at the output of the line. Today this is no longer true. As clock and data speeds continue to increase, seemingly without bound, these interconnect conductors will have a significant effect on the signal transmission and cannot be ignored.

— Dr. Clayton R. Paul

7.1 What Is a Transmission Line

A *transmission line* is a series of conductors, often but not necessarily two, used to guide electromagnetic energy from one place to the other [3]. It's that simple. The more complicated part is the math behind it (Maxwell's Equations; see Appendix E) because we no longer consider a transmission line a lumped-element RLC network (see Fig. 7.9). Rather more, we consider the signal conductor as a transmission line through which an electromagnetic field is moved from one point to another. Transmission lines are characterized or described by their characteristic impedance $\underline{Z}_0$ [Ω] and distributed parameter model (see Fig. 7.2).

Figure 7.1 shows some common transmission line geometries:

- **Coax.** The electromagnetic energy propagates through the dielectric (mostly PTFE, because of its low loss and stable ϵ_r' for many frequencies) between the center conductor and the inside surface of the outer conductor (shield) of a coaxial cable.
- **Microstrip.** Transmission line where the signal conductor is on the top or bottom layer of a PCB with an adjusted return path conductor (e.g., ground plane or power supply plane).
- **Stripline.** Transmission line where the signal conductor is embedded between two signal return path conductors of a PCB (e.g., ground or power supply plane).

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_7

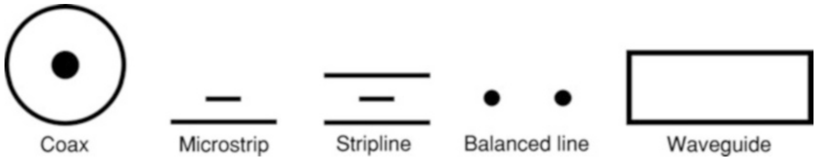


Fig. 7.1 Transmission line examples

- **Balanced line.** Two conductors of the same size and shape with equal impedance to ground and all other conductors (e.g., ethernet cable).
- **Waveguide.** A waveguide consists of a single hollow conductor used to guide the electromagnetic energy. Waveguides are used in the gigahertz frequency range, and they cannot pass DC signals.

7.2 When to Consider Transmission Lines

Every signal interconnection is a transmission line. However, it is not necessary to treat every signal path as a transmission line. Two different rules of thumb—if a conductor should be treated as transmission line or not—are explained in Sect. 7.2.1 (frequency-domain) and Sect. 7.2.2 (time-domain).

7.2.1 *Rule of Thumb for* $\mathbf{l_{critical}}$ *in the Frequency Domain*

A common rule of thumb when working in the frequency domain is [3]:

- **Interconnection length** $\mathbf{l \geq (\lambda_{min}/10)}$. Consider the signal path as a transmission line to minimize signal distortions and ringing due to reflections and to minimize radiated emissions and electromagnetic interference (EMI).
- **Interconnection length** $\mathbf{l < (\lambda_{min}/10)}$. Consider the signal path as a simple conductor. If an interconnection length l [m] is short with respect to the signal wavelength λ [m], it is good practice that the interconnection is considered as a simple conductor with lumped-element parameters (e.g., resistance R [Ω] with series inductance L [H]).

When determining the shortest wavelength λ_{min} [m] in a digital signal (e.g., clock), it is necessary to know the maximum frequency f_{max} [Hz] of the signal by considering the rising and falling time (rather than the fundamental frequency). The rule of thumb for calculating the bandwidth—or the highest significant sine wave frequency f_{max} [Hz]—of a rectangular digital signal and the corresponding minimum wavelength λ_{min} [m] [1]:

$$f_{max} = \text{bandwidth} = \frac{0.35}{t_{10-90\%}} \tag{7.1}$$

$$\lambda_{min} = \frac{v}{f_{max}} \tag{7.2}$$

where,

f_{max} = highest significant sine wave frequency harmonic in a digital signal (what significant means is described in Sect. 5.3) in [Hz]
$t_{10-90\%}$ = rise and/or fall time (whichever is smaller) from 10 % to 90 % of the slope of a digital signal in [sec]
λ_{min} = wavelength of the highest significant harmonic frequency in [m]
v = propagation velocity of the signal along the transmission line in [m/sec]

The frequency domain approximation for the critical length $l_{\text{critical,fd}}$ [m] can be calculated with respect to the digital signal rise /fall time $t_{10-90\%}$:

$$l_{\text{critical,fd}} = \frac{\lambda_{min}}{10} = \frac{v}{10 \cdot f_{max}} = \frac{v \cdot t_{10-90\%}}{10 \cdot 0.35} = \frac{c \cdot t_{10-90\%}}{3.5 \cdot \sqrt{\epsilon'_{r,eff}}} \tag{7.3}$$

where,

$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
v = propagation velocity of the signal along the signal line in [m/sec]
f_{max} = highest significant sine wave frequency harmonic in a digital signal (what significant means is described in Sect. 5.3) in [Hz]
$\epsilon'_{r,eff}$ = effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating

7.2.2 *Rule of Thumb for* $\mathbf{l_{critical}}$ *in the Time Domain*

There is also a rule of thumb for the time domain [3]:

- $\mathbf{t_{10-90\%} \leq 2 \cdot t_{pd}}$. If the rise or fall time $t_{10-90\%}$ [sec] of a digital signal is smaller than twice the propagation delay t_{pd} [sec] (along the signal line), then the signal path should be considered as a transmission line, in order to minimize signal distortions and ringing due to reflections and in order to minimize radiated emissions and electromagnetic interference (EMI).
- $\mathbf{t_{10-90\%} > 2 \cdot t_{pd}}$. If the rise and fall time $t_{10-90\%}$ [sec] of a digital signal are bigger than twice the propagation delay t_{pd} [sec] of the signal across the conductor, the signal path may be considered as a simple conductor.

If we multiply both sides of the equation of the time domain rule of thumb with the signal propagation velocity v [m/sec], we get:

$$t_{10-90\%} \leq 2 \cdot t_{pd}$$
$$t_{10-90\%} \cdot v \leq 2 \cdot t_{pd} \cdot v$$
$$l_{10-90\%} \leq 2 \cdot l_{\text{critical,td}}$$

where,

$t_{10-90\%}$ = rise and/or fall time (whichever is smaller) from 10 % to 90 % of the digital signal in [sec]
t_{pd} = propagation delay of the signal along the signal line in [sec]
$l_{10-90\%}$ = distance which the signal travels along the signal line during the rise and/or fall time (whichever is smaller) $t_{10-90\%}$ in [m]
$l_{\text{critical,td}}$ = critical length of the time domain rule of thumb in [m]

The time domain approximation for the critical length $l_{\text{critical,td}}$ [m] can be calculated with respect to the digital signal rise/fall time $t_{10-90\%}$:

$$l_{\text{critical,td}} = \frac{l_{10-90\%}}{2} = \frac{v \cdot t_{10-90\%}}{2} = \frac{c \cdot t_{10-90\%}}{2 \cdot \sqrt{\epsilon'_{r,eff}}} \tag{7.4}$$

where,

$l_{10-90\%}$ = distance which the signal travels along the signal line during the rise and/or fall time (whichever is smaller) $t_{10-90\%}$ in [m]
v = propagation velocity of the signal along the signal line in [m/sec]
$t_{10-90\%}$ = rise and/or fall time (whichever is smaller) from 10 to 90 % of the digital signal in [sec]
$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
$\epsilon'_{r,eff}$ = effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating

7.2.3 *Critical Length* $\mathbf{l_{critical}}$

The two rules above lead to similar values for the *critical length* l_{critical} [m], where the frequency domain $l_{\text{critical,fd}}$ is smaller than the time domain $l_{\text{critical,td}}$. It is therefore recommended to go with the rule of thumb of the frequency domain: Consider an interconnection as a transmission line if it is longer than:

$$l_{\text{critical}} = \frac{\lambda_{min}}{10} \tag{7.5}$$

where,

λ_{min} = wavelength of the highest significant harmonic frequency in the signal (what significant means is described in Sect. 5.3) in [m]

7.3 Characteristic Impedance Z_0

7.3.1 Characteristic Impedance Z_0 of Any Transmission Line

The *characteristic impedance* $\underline{Z}_0$ [Ω] is the most important property of a transmission line. The characteristic impedance $\underline{Z}_0$ [Ω] cannot be measured with a simple DC ohmmeter, and it is defined as the ratio of voltage $\underline{V}^+$ [V] to current $\underline{I}^+$ [A] of a *single* traveling wave along a transmission line [5]:

$$\underline{Z}_0 = \frac{\underline{V}^+}{\underline{I}^+} \tag{7.6}$$

where

$\underline{V}^+$ = voltage of the forward wave (the $^+$ emphasizes that voltage and current travel in the same direction) in complex phasor form in [V]

$\underline{I}^+$ = current of the forward wave (the $^+$ emphasizes that voltage and current travel in the same direction) in complex phasor form in [A]

Figure 7.2 shows the equivalent circuit model for a transmission line with two conductors. The so-called distributed parameter model divides the transmission line into infinitely small segments of length dz in [m]. The parameters of these segments are defined per-unit-length (e.g., per [m]):

$$R' = \text{Resistance per-unit-length } [\Omega/\text{m}] \tag{7.7}$$

$$L' = \text{Inductance per-unit-length } [\text{H}/\text{m}] \tag{7.8}$$

$$C' = \text{Capacitance per-unit-length } [\text{F}/\text{m}] \tag{7.9}$$

$$G' = \text{Conductance per-unit-length } [\text{S}/\text{m}] \tag{7.10}$$

The formula for the characteristic impedance $\underline{Z}_0$ [Ω] of a transmission line—based on the distributed parameter model—is defined as [3]:

$$\underline{Z}_0 = \sqrt{\frac{R' + j\omega L'}{G' + j\omega C'}} \tag{7.11}$$

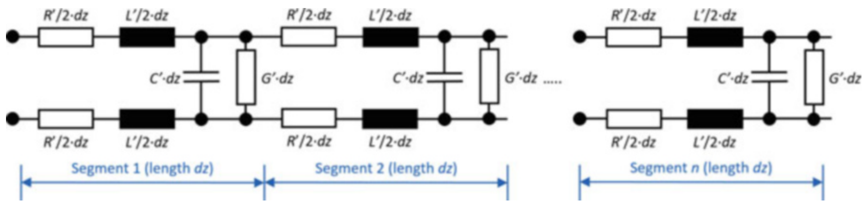


Fig. 7.2 Distributed parameter model of a transmission line [3]

where,

$\omega = 2\pi f$ = angular frequency in [rad/sec]
$j = \sqrt{-1}$ = imaginary unit

The following subsections present formulas for the calculation of the characteristic impedance $\underline{Z}_0$ [Ω] for typical electrical transmission lines (the formulas are approximations; in case you need accurate results, you should use a solver that applies Maxwell's Equations presented in Appendix E):

- **Lossless transmission lines:** Sect. 7.3.2.
- **Parallel wires:** Sect. 7.3.3.
- **Twisted pairs:** Sect. 7.3.4.
- **Coaxial cables:** Sect. 7.3.5.
- **Microstrip lines:** Sect. 7.3.6.
- **Coplanar waveguide with reference plane:** Sect. 7.3.7.
- **Centered striplines:** Sect. 7.3.8.

7.3.2 Characteristic Impedance Z_0 of Lossless Transmission Lines

In practice, it is often adequate to describe transmission lines as lossless ($R' = 0$, $G' = 0$). In that case, the transmission line model can be simplified—like shown in Fig. 7.3—and the calculation of the characteristic impedance is reduced to:

$$Z_0 = \sqrt{\frac{L'}{C'}} \tag{7.12}$$

where,

L' = inductance per-unit-length [H/m]
C' = capacitance per-unit-length [F/m]

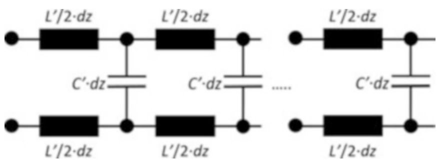
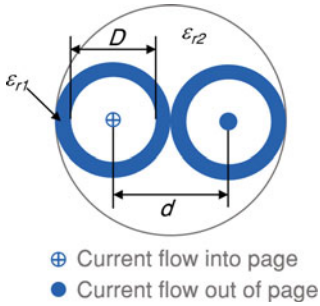


Fig. 7.3 Distributed parameter model of a loss less transmission line [3]

Fig. 7.4 Cross section of two parallel wires or twisted pair



7.3.3 Characteristic Impedance Z_0 of Parallel Wires

The characteristic impedance $\underline{Z}_0$ [Ω] of two parallel wires with identical diameter D [m]—according to Fig. 7.4—can be calculated as [6]:

$$Z_0 = \frac{\eta_0}{\pi\sqrt{\epsilon'_{r,eff}}}\cosh^{-1}\left(\frac{d}{D}\right) \tag{7.13}$$

$$\epsilon'_{r,eff} = \epsilon'_{r2} + 0.25\left(\epsilon'_{r1} - \epsilon'_{r2}\right) \tag{7.14}$$

where,

$\eta_0 = 377\,\Omega$ = intrinsic impedance of free-space (vacuum) [Ω]
d = distance from center to center of the two wires in [m]
D = diameter of the two wires in [m]
$\epsilon'_{r,eff}$ = effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating
ϵ'_{r1} = relative permeability of the insulation around the conductors
ϵ'_{r2} = relative permeability of the medium around the insulated conductors

7.3.4 Characteristic Impedance Z_0 of Twisted Pairs

The characteristic impedance $\underline{Z}_0$ [Ω] of a twisted pair with identical diameter D [m]—according to Fig. 7.4—can be calculated as [6]:

$$Z_0 = \frac{\eta_0}{\pi\sqrt{\epsilon'_{r,eff}}}\cosh^{-1}\left(\frac{d}{D}\right) \tag{7.15}$$

$$\epsilon'_{r,eff} = \epsilon'_{r2} + q\left(\epsilon'_{r1} - \epsilon'_{r2}\right) \tag{7.16}$$

$$q = 0.25 + 0.0004\theta \tag{7.17}$$

$$T = \frac{\tan(\theta)}{\pi d} \tag{7.18}$$

$$\theta = \tan^{-1}(T\pi d) \tag{7.19}$$

where,

$\eta_0 = 377\,\Omega$ = intrinsic impedance of free space (vacuum) [Ω]
d = distance from center to center of the two wires in [m]
D = diameter of the two wires in [m]
$\epsilon'_{r,eff}$ = effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating
ϵ'_{r1} = relative permeability of the insulation around the conductors
ϵ'_{r2} = relative permeability of the medium around the insulated conductors
θ = angle between the twisted pair's center line and the twist in [rad]
T = twists per length in [1/m]

7.3.5 Characteristic Impedance Z_0 of Coaxial Cables

The characteristic impedance $\underline{Z}_0$ [Ω] of a round coaxial cable—according to Fig. 7.5—can be calculated as [6]:

$$Z_0 = \frac{\eta_0}{2\pi\sqrt{\epsilon'_r}} \ln\left(\frac{D_{coax}}{D_{core}}\right) \tag{7.20}$$

where,

$\eta_0 = 377\,\Omega$ = intrinsic impedance of free space (vacuum) [Ω]
D_{coax} = inner diameter of the coaxial cable shield in [m]
D_{core} = outer diameter of the core wire of the coaxial cable in [m]
ϵ'_r = dielectric constant of the dielectric between core and shield

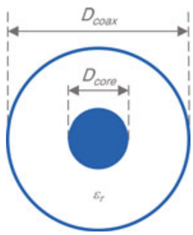


Fig. 7.5 Coaxial cable

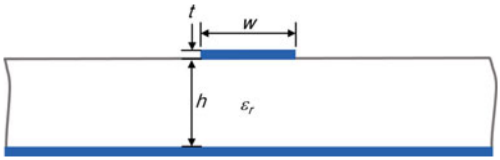


Fig. 7.6 Microstrip line

7.3.6 Characteristic Impedance Z_0 of Microstrip Lines

The characteristic impedance $\underline{Z}_0$ [Ω] of a microstrip line—according to Fig. 7.6—can be calculated as [6]:

$$Z_0 = \frac{\eta_0}{2\pi\sqrt{2}\sqrt{\epsilon_r' + 1}} \tag{7.21}$$

$$\ln\left(1 + \frac{4h}{w'}\left(\frac{14 + 8/\epsilon_r'}{11}\frac{4h}{w'} + \sqrt{\left(\frac{14 + 8/\epsilon_r'}{11}\right)^2\left(\frac{4h}{w'}\right)^2 + \frac{1 + 1/\epsilon_r'}{2}\pi^2}\right)\right) \tag{7.22}$$

$$w' = w + \Delta w' \tag{7.23}$$

$$\Delta w' = \Delta w\left(\frac{1 + 1/\epsilon_r'}{2}\right) \tag{7.24}$$

$$\Delta w = \frac{t}{\pi}\ln\left(\frac{4e}{\sqrt{\left(\frac{t}{h}\right)^2 + \left(\frac{1/\pi}{w/t+1.1}\right)^2}}\right) \tag{7.25}$$

where,

$\eta_0 = 377\,\Omega$ = intrinsic impedance of free space (vacuum) [Ω]
h = distance of the PCB trace above the reference (ground) plane in [m]
w = width of the PCB trace in [m]
w' = corrected PCB trace width due to thickness t in [m]
t = PCB trace thickness in [m]
ϵ_r' = dielectric constant of the PCB dielectric material

7.3.7 Characteristic Impedance Z_0 of Coplanar Waveguide with Reference Plane

The characteristic impedance $\underline{Z}_0$ [Ω] of a coplanar waveguide with reference plane (ground) according to Fig. 7.7 can be calculated as [6]:

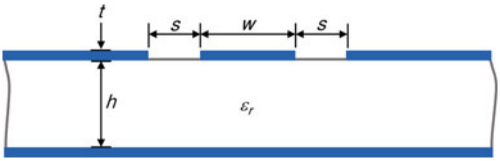


Fig. 7.7 Coplanar waveguide with reference plane (ground) or microstrip line with signal side reference plane

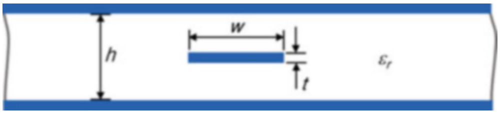


Fig. 7.8 Stripline

$$Z_0 = \frac{\pi 60\,\Omega}{\sqrt{\epsilon'_{r,eff}}} \frac{1}{\frac{K(k)}{K(k')} + \frac{K(k_1)}{K(k'_1)}} \tag{7.26}$$

$$k = \frac{w}{w + 2s} \tag{7.27}$$

$$k_1 = \frac{\tanh\left(\frac{\pi w}{4h}\right)}{\tanh\left(\frac{\pi(w+2s)}{4h}\right)} \tag{7.28}$$

$$k' = \sqrt{1 - k^2} \tag{7.29}$$

$$k'_1 = \sqrt{1 - k_1^2} \tag{7.30}$$

$$\epsilon'_{r,eff} = \frac{1 + \epsilon_r \frac{K(k')}{K(k)} \frac{K(k_1)}{K(k'_1)}}{1 + \frac{K(k')}{K(k)} \frac{K(k_1)}{K(k'_1)}} \tag{7.31}$$

where,

$K(x)$ = elliptic integral of the first kind of x (x stands for k, k', k_1, k'_1)
h = distance between the two reference (ground) planes in [m]
w = width of the PCB trace in [m]
s = space between the PCB trace and the coplanar reference plane in [m]
$\epsilon'_{r,eff}$ = effective dielectric constant of the PCB dielectric material

7.3.8 Characteristic Impedance Z_0 of Centered Striplines

The characteristic impedance $\underline{Z}_0$ [Ω] of a stripline—which is located in the middle of two reference planes according to Fig. 7.8—can be calculated as [6]:

$$Z_0 = \frac{60\,\Omega}{\sqrt{\epsilon_r'}} \ln\left(\frac{4h}{\pi K_1}\right), \text{ for } \frac{w}{h} \leq 0.35 \tag{7.32}$$

$$Z_0 = \frac{94.15\,\Omega}{\frac{w/h}{1-t/h} + \frac{K_2}{\pi}} \frac{1}{\sqrt{\epsilon_r'}}, \text{ for } \frac{w}{h} > 0.35 \tag{7.33}$$

$$K_1 = \frac{w}{2}\left[1 + \frac{t}{\pi w}\left(1 + \ln\left(\frac{4\pi w}{t}\right) + 0.51\pi\left(\frac{t}{w}\right)^2\right)\right] \tag{7.34}$$

$$K_2 = \frac{2}{1-t/h} \ln\left(\frac{1}{1-t/h} + 1\right) - \left(\frac{1}{1-t/h} - 1\right) \ln\left(\frac{1}{(1-t/h)^2} - 1\right) \tag{7.35}$$

where,

h = distance between the two reference (ground) planes in [m]
t = thickness of the PCB trace in [m]
w = width of the PCB trace in [m]
ϵ_r' = dielectric constant of the PCB dielectric material

7.4 Per-Unit-Length Inductance L′ and Capacitance C′

Electrically short transmission lines ($l < \lambda/10$) can be modeled as lumped-element RLC networks (Fig. 7.9). The following subsections present formulas for the calculation of the *per-unit-length inductance* L' [H/m] and the *per-unit-length capacitance* C' [F/m] for typical electrical transmission lines:

- **Two-wire lines:** Sect. 7.4.1.
- **Wires above ground plane:** Sect. 7.4.2.
- **Coaxial cables:** Sect. 7.4.3.
- **PCB traces:** Sect. 7.4.4.

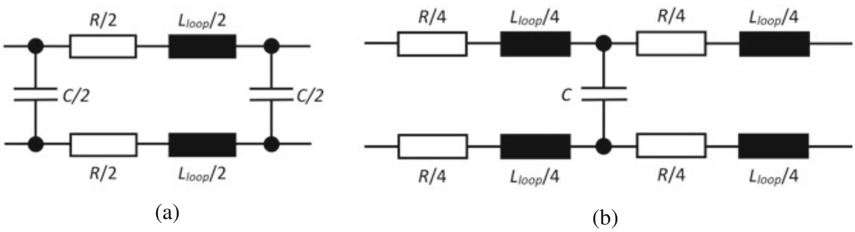


Fig. 7.9 Lumped-element RLC-network examples of transmission lines. (**a**) Lumped π-model. (**b**) Lumped T-model

7.4.1 *L′ and C′ of Two-Wire Lines*

For a setup with two parallel wires (e.g., ribbon cable with identical diameter D [m]), where the current flows in the opposite direction, the per-unit-length inductance L' [H/m] and the per-unit-length capacitance C' [F/m] are given as [4]:

$$L'_{TwoWireLine} \approx \frac{\mu'_{r,eff}\mu_0}{\pi} \cosh^{-1}\left(\frac{d}{D}\right) \tag{7.36}$$

$$C'_{TwoWireLine} \approx \frac{\pi \epsilon'_{r,eff}\epsilon_0}{\cosh^{-1}\left(\frac{d}{D}\right)} \tag{7.37}$$

where,

$\cosh^{-1}$ = area hyperbolic cosine (inverse hyperbolic cosine) function [1, $+\infty$]
$\mu'_{r,eff}$ = effective relative permeability of the environment of the two wires and therefore the material through which the magnetic flux flows
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon'_{r,eff}$ = effective relative permittivity (dielectric constant) of the environment of the two wires and, therefore, the material through which the electric field lines are formed
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity
d = distance from center to center of the two wires in [m]
D = diameter of the two wires in [m]

The effective relative permeability $\mu'_{r,eff}$ is often equal 1.0, because the insulation material of cables and the environment around the lines is usually non-magnetic (e.g., plastics, aluminum, or copper).

The calculation of the effective relative permeability $\epsilon'_{r,eff}$ is often complicated because there are different dielectric materials involved (e.g., cable insulation and air). The approximate effective dielectric constant $\epsilon'_{r,eff}$ for a pair of wire—like shown in Fig. 7.10—is given in as [6]:

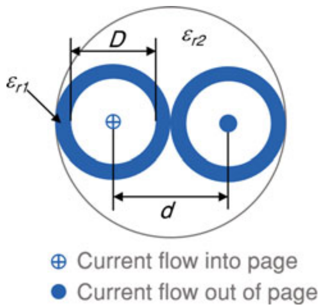


Fig. 7.10 Two-wire line

Table 7.1 Unit-per-length inductance L' and capacitance C' of two-wire lines with PVC insulation around each wire

AWG	d [m]	μ_{reff} [1]	ε_{r1} [1]	ε_{r2} [1]	ε_{reff} [1]	L'_{loop} [H/m]	C' [F/m]
AWG6	0.0081	1	3	1	1.5	0.52µH/m	32.1pF/m
AWG20	0.0026	1	3	1	1.5	0.73µH/m	22.7pF/m
AWG32	0.00075	1	3	1	1.5	0.80µH/m	21.0pF/m

(a) PVC insulates wires, side-by-side without cable jacket.

AWG	d [m]	μ_{reff} [1]	ε_{r1} [1]	ε_{r2} [1]	ε_{reff} [1]	L'_{loop} [H/m]	C' [F/m]
AWG6	0.0081	1	3	3	3	0.52µH/m	64.2pF/m
AWG20	0.0026	1	3	3	3	0.73µH/m	45.5pF/m
AWG32	0.00075	1	3	3	3	0.80µH/m	42.0pF/m

(b) PVC insulates wires, side-by-side with PVC cable jacket.

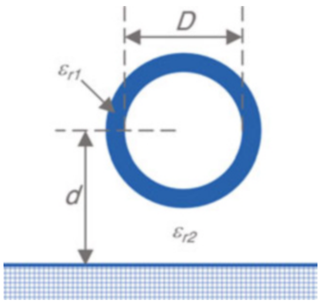


Fig. 7.11 Wire above ground plane

$$\epsilon'_{r,eff} \approx \epsilon'_{r2} + 0.25\left(\epsilon'_{r1} - \epsilon'_{r2}\right) \tag{7.38}$$

where,

ϵ'_{r1} = relative permeability of the insulation around the conductors
ϵ'_{r2} = relative permeability of the medium around the insulated conductors

Based on Eqs. 7.36 and 7.37, Table 7.1a lists examples of per-unit-length inductance L' and per-unit-length capacitance C' of a two-wire line without an additional cable jacket ($\epsilon_{r2} = 1$). On the other hand, Table 7.1b lists the values for L' and C' with a two-wire line and a PVC cable jacket around the wires ($\epsilon_{r2} = 3$).

7.4.2 L' and C' of Wires Above Ground Plane

In the case of a wire above an infinite ground plane (Fig. 7.11), the per-unit-length loop inductance L' [H/m] and the per-unit-length capacitance C' [F/m] are given as [4]:

$$L'_{WireAboveGndPlane} \approx \frac{\mu'_{r,eff}\mu_0}{2\pi}\cosh^{-1}\left(\frac{2d}{D}\right) \tag{7.39}$$

$$C'_{WireAboveGndPlane} \approx \frac{2\pi\epsilon'_{r,eff}\epsilon_0}{\cosh^{-1}\left(\frac{2d}{D}\right)} \tag{7.40}$$

where,

$\cosh^{-1}$ = area hyperbolic cosine (inverse hyperbolic cosine) function $[1, +\infty]$
$\mu'_{r,eff}$ = effective relative permeability of the environment of the wire and therefore the material through which the magnetic flux flows
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon'_{r,eff}$ = effective relative permittivity (dielectric constant) between the wire and the ground plane and therefore the material through which the electric field lines are formed
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity
d = distance from the center of the wire to the surface of the ground plane [m]
D = diameter of the wire in [m]

The effective relative permeability $\mu'_{r,eff}$ is usually equal 1.0, because all the media around the conductor are non-magnetic (e.g. plastics, copper), whereas the effective relative permittivity $\epsilon'_{r,eff}$ between the wire and the ground plane is very difficult to calculate, as it depends very much on the distance of the wire to the ground plane, the thickness of the insulation around the wire, and the media between the wire and the ground plane. Table 7.2 gives an idea about the ranges of the per-unit-length inductance L' [H/m] and the per-unit-capacitance C' [F/m] of a single wire above a ground plane in case the wire is close to the ground plane or far away.

Table 7.2 Unit-per-length inductance L' and capacitance C' of a single PVC insulated wire above an infinite ground plane

AWG	d [m]	μ_{reff} [1]	ε_{r1} [1]	ε_{r2} [1]	ε_{reff} [1]	L'_{loop} [H/m]	C' [F/m]
AWG6	0.00405	1	3	1	1.5	0.26μH/m	64.2pF/m
AWG20	0.0013	1	3	1	1.5	0.36μH/m	45.5pF/m
AWG32	0.00038	1	3	1	1.5	0.40μH/m	42.0pF/m

(a) PVC insulated wire laying directly on ground plane.

AWG	d [m]	μ_{reff} [1]	ε_{r1} [1]	ε_{r2} [1]	ε_{reff} [1]	L'_{loop} [H/m]	C' [F/m]
AWG6	0.01	1	1	1	1	0.45μH/m	24.6pF/m
AWG20	0.01	1	1	1	1	0.78μH/m	14.3pF/m
AWG32	0.01	1	1	1	1	1.06μH/m	10.1pF/m

(b) PVC insulated wire 10mm above ground plane.

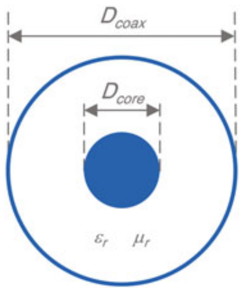


Fig. 7.12 Cross section of a coaxial cable

Table 7.3 Unit-per-length inductance L' and capacitance C' of different coaxial cables

RG	D_{coax} [m]	D_{core} [m]	μ_r [1]	ε_r [1]	L'_{loop} [H/m]	C' [F/m]	Z_0 [Ω]
RG58	0.003	0.0009	1	2.1	0.24μH/m	97pF/m	50
RG11	0.0073	0.0012	1	2.1	0.36μH/m	64.7pF/m	75
RG6	0.0046	0.00072	1	2.2	0.37μH/m	66.0pF/m	75

7.4.3 L′ and C′ of Coaxial Cables

For a coaxial cable (Fig. 7.12), the per-unit-length loop inductance L' [H/m] and the per-unit-length capacitance C' [F/m] are given as [4]:

$$L'_{coax} \approx \frac{\mu_r \mu_0}{2\pi} \ln\left(\frac{D_{coax}}{D_{core}}\right) \tag{7.41}$$

$$C'_{coax} \approx \frac{2\pi\epsilon_r\epsilon_0}{\ln\left(\frac{D_{coax}}{D_{core}}\right)} \tag{7.42}$$

where,

μ_r = relative permeability of the coaxial cable dielectric material in [1]
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
ϵ'_r = relative permittivity of the coaxial cable dielectric material in [1]
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity
D_{coax} = inner diameter of the coaxial cable shield in [m]
D_{core} = outer diameter of the core wire of the coaxial cable in [m]

Table 7.3 gives an idea about the ranges of the per-unit-length inductance L' [H/m] and the per-unit-capacitance C' [F/m] of typical coaxial cables.

7.4.4 *L′ and C′ of PCB Traces*

It is not common to specify the per-unit-length inductance L' [H/m] and the per-unit-length capacitance C' [F/m] for PCB traces. Instead, it is more common to specify the characteristic impedance Z_0 [Ω] of the transmission line (PCB trace), which can be expressed as:

$$Z_0 = \sqrt{\frac{L'}{C'}} \tag{7.43}$$

Another fundamental parameter is the propagation velocity v [m/sec] of a signal along a transmission line:

$$v = \frac{1}{\sqrt{L'C'}} = \frac{c}{\sqrt{\mu'_{r,eff}\epsilon'_{r,eff}}} \tag{7.44}$$

$$v = \frac{c}{\sqrt{\epsilon'_{r,eff}}}, \text{ because usually } \mu'_{r,eff} = 1.0 \tag{7.45}$$

Thus, we can write:

$$L' = \frac{Z_0}{v} = \frac{Z_0\sqrt{\epsilon'_{r,eff}}}{c} \tag{7.46}$$

$$C' = \frac{1}{vZ_0} = \frac{\sqrt{\epsilon'_{r,eff}}}{cZ_0} \tag{7.47}$$

where,

L' = per-unit-length inductance of the transmission line (e.g., PCB trace) [H/m]
C' = per-unit-length capacitance of the transmission line (e.g., PCB trace) [F/m]
$c = 1/(\sqrt{\mu_0\epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
$\mu'_{r,eff}$ = effective magnetic relative permeability through which the electromagnetic wave is propagating
$\epsilon'_{r,eff}$ = effective dielectric constant (relative permittivity) through which the electromagnetic wave is propagating

Tables 7.4, 7.5 and, 7.6 present some example values of per-unit-length inductance L' [H/m] and per-unit-length capacitance C' [F/m] for microstrip lines, striplines, and coplanar waveguides with reference plane. All values are approximations.

Formulas for calculating the characteristic impedance Z_0 of some selected PCB structures can be found in the next Chap. 11.1.7.2.

Table 7.4 Unit-per-length inductance L' and capacitance C' of mirostrip lines shown in Fig. 7.6. t = thickness of the PCB trace. h = distance of the PCB trace above the reference plane. w = width of the PCB trace

t [m]	h [m]	w [m]	ε_r [1]	$\varepsilon_{r,eff}$ [1]	L'_{loop} [H/m]	C' [F/m]	Z_0 [Ω]
3.50E−05	0.0015	0.0001	4.5	2.9	0.90μH/m	33.6pF/m	157
3.50E−05	0.0015	0.001	4.5	3.2	0.49μH/m	71.4pF/m	83
3.50E−05	0.0015	0.01	4.5	3.8	0.13μH/m	320pF/m	20
3.50E−05	0.0001	0.0001	4.5	3.2	0.38μH/m	94.6pF/m	63
3.50E−05	0.0001	0.001	4.5	3.9	0.09μH/m	462pF/m	14
3.50E−05	0.0001	0.01	4.5	4.4	0.01μH/m	4040pF/m	1.7

Table 7.5 Unit-per-length inductance L' and capacitance C' of coplanar waveguides with a reference plane shown in Fig. 7.7. h = distance between the two reference (ground) planes. w = width of the PCB trace. s = space between the PCB trace and the coplanar reference plane

h[m]	w [m]	s [m]	ε_r [1]	$\varepsilon_{r,eff}$ [1]	L'_{loop} [H/m]	C' [F/m]	Z_0 [Ω]
0.0015	0.0001	0.001	4.5	2.82	0.85μH/m	36.8pF/m	152.0
0.0015	0.001	0.001	4.5	2.95	0.43μH/m	75.3pF/m	76.0
0.0015	0.01	0.001	4.5	3.66	0.13μH/m	325pF/m	19.6
0.0001	0.0001	0.001	4.5	3.5	0.48μH/m	80.5pF/m	77.5
0.0001	0.001	0.001	4.5	4.1	0.10μH/m	450pF/m	15.0
0.0001	0.01	0.001	4.5	4.47	0.005μH/m	10'010pF/m	0.7
0.0015	0.0001	0.0001	4.5	2.75	0.43μH/m	62.1pF/m	89.0
0.0015	0.001	0.0001	4.5	2.8	0.25μH/m	123pF/m	45.5
0.0015	0.01	0.0001	4.5	3.42	0.10μH/m	376pF/m	16.4
0.0001	0.0001	0.0001	4.5	3.09	0.39μH/m	87.5pF/m	67.0
0.0001	0.001	0.0001	4.5	3.87	0.09μH/m	455pF/m	14.4
0.0001	0.01	0.0001	4.5	4.44	0.005μH/m	10'018pF/m	0.7

Table 7.6 Unit-per-length inductance L' and capacitance C' of striplines shown in Fig. 7.8. t = thickness of the PCB trace. h = distance between the two reference planes. w = width of the PCB trace

t [m]	h [m]	w [m]	ε_r [1]	L'_{loop} [H/m]	C' [F/m]	Z_0 [Ω]
3.50E−05	0.0015	0.0001	4.5	0.64μH/m	77.7pF/m	91.0
3.50E−05	0.0015	0.001	4.5	0.27μH/m	186pF/m	38.1
3.50E−05	0.0015	0.01	4.5	0.043μH/m	1164pF/m	6.1
3.50E−05	0.0001	0.0001	4.5	0.13μH/m	382pF/m	18.5
3.50E−05	0.0001	0.001	4.5	0.019μH/m	2588pF/m	2.7
3.50E−05	0.0001	0.01	4.5	0.002μH/m	24'650pF/m	0.3

7.5 Propagation Constant γ

The *propagation constant* $\underline{\gamma}$ [1/m]—also called propagation factor—describes the attenuation and phase shift of the signal as it propagates through the transmission line. To be even more precise: the propagation constant of a sinusoidal electromagnetic wave is a measure of the change undergone by the amplitude and phase of the wave as it propagates in a given direction (Fig. 7.13).

Let's imagine a sinusoidal voltage, current, electric field, or magnetic field which propagates in the direction of z [m] and which has amplitude of $\underline{A}_0$ at its source and an amplitude of $\underline{A}_z$ at the distance z [m] from the source. Then $\underline{A}_z$ can be written as:

$$\underline{A}_z(z) = \underline{A}_0 \cdot e^{-\underline{\gamma} z} \tag{7.48}$$

where,

$\underline{A}_z$ = complex phasor of a sinusoidal voltage [V], current [A], electric field [V/m], or magnetic field [A/m] at the distance z away from the source
$\underline{A}_0$ = complex phasor of a sinusoidal voltage [V], current [A], electric field [V/m], or magnetic field [A/m] at the source
$\underline{\gamma}$ = the complex propagation constant in [1/m]
z = distance traveled along a transmission line in [m]

The propagation constant $\underline{\gamma}$ [1/m] can be calculated based on the per-unit-length parameters of a transmission line [2]:

$$\underline{\gamma} = \sqrt{(R' + j\omega L')(G' + j\omega C')} = j\omega\sqrt{\underline{\epsilon\mu}} \tag{7.49}$$

where,

R' = resistance per-unit-length in [Ω/m]
L' = inductance per-unit-length in [H/m]
C' = capacitance per-unit-length in [F/m]

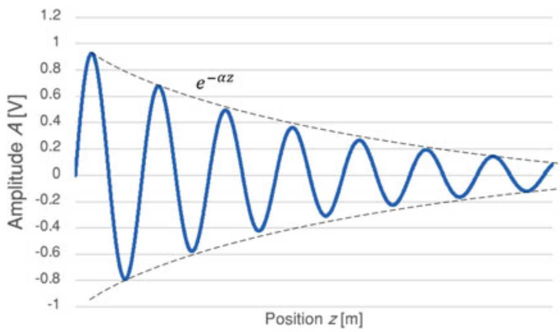


Fig. 7.13 Attenuation of a sinusoidal signal along a transmission line with attenuation factor α [1/m]

G' = conductance per-unit-length in [S/m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\underline{\mu} = \mu' - j\mu''$ = complex permeability of the media through which the electromagnetic wave is propagating in [H/m]
$\underline{\epsilon} = \epsilon' - j\epsilon''$ = complex permittivity of the media through which the electromagnetic wave is propagating in [F/m]

More details about the complex permittivity and permeability can be found in Appendix Sects. D.5 and D.6.

Another notation of $\underline{\gamma}$ is given as [2]:

$$\underline{\gamma} = \alpha + j\beta \tag{7.50}$$

where,

α = attenuation constant (or attenuation factor) in [1/m]
β = phase constant (or phase factor) in [rad/m]

This means that the transmission line Eq. 7.48 can be rewritten as:

$$\underline{A}_z = \underline{A}_0 \cdot e^{-\alpha z} e^{-j\beta z} \tag{7.51}$$

In a general form, the *attenuation constant* α [1/m] and the *phase constant* β [rad/m] can be calculated like this [2]:

$$\alpha = \omega \sqrt{\frac{(\epsilon'\mu' - \epsilon''\mu'')}{2} \cdot \left(\sqrt{1 + \left(\frac{\epsilon'\mu'' + \epsilon''\mu'}{\epsilon'\mu' - \epsilon''\mu''} \right)^2} - 1 \right)} \tag{7.52}$$

$$\beta = \frac{2\pi}{\lambda} = \omega \sqrt{\frac{(\epsilon'\mu' - \epsilon''\mu'')}{2} \cdot \left(\sqrt{1 + \left(\frac{\epsilon'\mu'' + \epsilon''\mu'}{\epsilon'\mu' - \epsilon''\mu''} \right)^2} + 1 \right)} \tag{7.53}$$

where,

$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
ϵ' = real part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
ϵ'' = imaginary part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
μ' = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]
μ'' = imaginary part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]

The magnetic loss μ'' can often be neglected ($\mu'' = 0$) and the dielectric loss ϵ'' can be written as [6]:

$$\underline{\epsilon} = \epsilon' - j\epsilon'' = \epsilon' - j\frac{\sigma}{\omega} \tag{7.54}$$

$$\epsilon'' = \frac{\sigma}{\omega} \tag{7.55}$$

where,

ϵ'' = dielectric loss = energy dissipated when a medium is under the influence of an external electric field = imaginary part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
σ = specific conductance of the medium where the wave is propagating through in [S/m]

With $\mu'' = 0$ and $\epsilon'' = \sigma/\omega$, the attenuation constant α [1/m] and the phase constant β [rad/m] can be calculated like this [6]:

$$\alpha = \omega \sqrt{\frac{\mu'\epsilon'}{2} \cdot \left(\sqrt{1 + \left(\frac{\sigma}{\omega\epsilon'} \right)^2} - 1 \right)} \tag{7.56}$$

$$\beta = \omega \sqrt{\frac{\mu'\epsilon'}{2} \cdot \left(\sqrt{1 + \left(\frac{\sigma}{\omega\epsilon'} \right)^2} + 1 \right)} \tag{7.57}$$

For a lossless line ($\sigma = 0$) we get [2]:

- No attenuation α [1/m]:

$$\alpha = 0 \tag{7.58}$$

- Phase shift β [rad/m]:

$$\beta = \omega\sqrt{R'C'} = \omega\sqrt{\epsilon'\mu'} \tag{7.59}$$

where,

$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\epsilon' = \epsilon_r'\epsilon_0$ = ability to store energy in a medium when an external electric field is applied = real part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
$\mu' = \mu_r\mu_0$ = ability to store energy in a medium when an external magnetic field is applied = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]

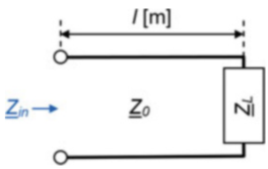


Fig. 7.14 Transmission line input impedance $\underline{Z}_{in}(l)$

7.6 Input Impedance of Transmission Lines

7.6.1 Input Impedance of Any Transmission Line

The *characteristic impedance* $\underline{Z}_0$ [Ω] of a transmission line—as already discussed in Chap. 7.3—is the ratio of the amplitude of a *single* voltage wave to its current wave. Since most transmission lines also have a reflected wave, the characteristic impedance is generally not the impedance that is measured on the line. The impedance $\underline{Z}_{in}(l)$ [Ω] (Fig. 7.14) measured at a given distance l [m] from the load impedance $\underline{Z}_L$ [Ω] can be expressed as [6]:

$$\underline{Z}_{in}(l) = \frac{\underline{V}(l)}{\underline{I}(l)} = \underline{Z}_0 \frac{\underline{Z}_L + \underline{Z}_0 \tanh\left(\underline{\gamma} l\right)}{\underline{Z}_0 + \underline{Z}_L \tanh\left(\underline{\gamma} l\right)} = \underline{Z}_0 \frac{\underline{Z}_L \cosh\left(\underline{\gamma} l\right) + \underline{Z}_0 \sinh\left(\underline{\gamma} l\right)}{\underline{Z}_0 \cosh\left(\underline{\gamma} l\right) + \underline{Z}_L \sinh\left(\underline{\gamma} l\right)} \tag{7.60}$$

where,

$\underline{Z}_{in}(l)$ = impedance measured at distance l from the load $\underline{Z}_L$ in [Ω]
$\underline{V}(l)$ = voltage at distance l from the load $\underline{Z}_L$ in [V]
$\underline{I}(l)$ = current at distance l from the load $\underline{Z}_L$ in [V]
$\underline{Z}_0$ = the complex characteristic impedance of the transmission line in [Ω]
$\underline{Z}_L$ = the complex load impedance in [Ω]
$\underline{\gamma} = \alpha + j\beta$ = the complex propagation constant in [1/m]
l = the distance from where the input of the transmission line to the load in [m]

7.6.2 Input Impedance of a Lossless Transmission Line

For a *lossless transmission line*, the propagation constant is purely imaginary [6]:

$$\alpha = 0 \tag{7.61}$$

$$\underline{\gamma} = j\beta = \frac{2\pi j}{\lambda} \tag{7.62}$$

and the input impedance to a lossless transmission line can be calculated as:

$$\underline{Z}_{in}(l) = \underline{Z}_0 \frac{\underline{Z}_L + j\underline{Z}_0 \tan(\beta l)}{\underline{Z}_0 + j\underline{Z}_L \tan(\beta l)} = \underline{Z}_0 \frac{\underline{Z}_L \cos(\beta l) + j\underline{Z}_0 \sin(\beta l)}{\underline{Z}_0 \cos(\beta l) + j\underline{Z}_L \sin(\beta l)} \tag{7.63}$$

where,

$\underline{Z}_{in}(l)$ = impedance measured at distance l from the load $\underline{Z}_L$ in [Ω]
$\underline{Z}_0$ = the complex characteristic impedance of the transmission line in [Ω]
$\underline{Z}_L$ = the complex load impedance in [Ω]
l = the distance from where the input of the transmission line to the load in [m]
$\beta = 2\pi/\lambda$ = phase constant (or phase factor) in [rad/m]

7.6.3 *Input Impedance of a Transmission Line At* $l = \lambda/2$

When the distance from the input of the transmission line to the load is a multiple of $\lambda/2$ ($l = n\lambda/2$) and therefore $\beta l = n\pi$ (where n is an integer), the input impedance to the transmission line $\underline{Z}_{in}(l)$ is equal the load impedance $\underline{Z}_L$ [6]:

$$\underline{Z}_{in}(l) = \underline{Z}_L \tag{7.64}$$

7.6.4 *Impedance of a Transmission Line At* $l = \lambda/4$

When the distance from the input of the transmission line to the load is a multiple of $\lambda/4$ ($\beta l = n\pi/2$) and therefore $l = n\lambda/4$ (where n is an integer), the input impedance to the transmission line $\underline{Z}_{in}(l)$ is [6]:

$$\underline{Z}_{in}(l) = \frac{\underline{Z}_0^2}{\underline{Z}_L} \tag{7.65}$$

This means in case $\underline{Z}_L = 0$, the transmission line input impedance becomes $\underline{Z}_{in} = \infty$ and vice versa.

7.6.5 *Impedance of a Matched Transmission Line*

In case the load impedance $\underline{Z}_L$ is equal the characteristic impedance $\underline{Z}_0$ of the transmission line (in other words: the load is matched), we can write:

$$\underline{Z}_{in}(l) = \underline{Z}_0 = \underline{Z}_L \tag{7.66}$$

7.6.6 *Impedance of a Shorted Transmission Line*

For the case of a shorted load $\underline{Z}_L = 0$, the input impedance $\underline{Z}_{in}(l)$ is purely imaginary and a periodic function of position and wavelength λ [m] (frequency f [Hz]):

$$\underline{Z}_{in}(l) = j\underline{Z}_0 \tan(\beta l) = j\underline{Z}_0 \tan\left(\frac{2\pi l}{\lambda}\right) \tag{7.67}$$

7.6.7 *Impedance of an Open Transmission Line*

For the case of an open load $\underline{Z}_L = \infty$, the input impedance $\underline{Z}_{in}(l)$ is purely imaginary and a periodic function of position and wavelength λ [m] (frequency f [Hz])

$$\underline{Z}_{in}(l) = -j\underline{Z}_0 \cot(\beta l) = -j\underline{Z}_0 \cot\left(\frac{2\pi l}{\lambda}\right) \tag{7.68}$$

7.7 High-Frequency Losses

The lossless line model is often accurate enough for frequencies up 100 MHz. However, above 100 MHz, the *high-frequency losses* may not be neglected anymore, and therefore the attenuation factor α [1/m] cannot be assumed to be zero. High-frequency losses are a result of [3]:

- **Ohmic Loss.** *Ohmic losses* result from the resistance of the conductors. Ohmic losses increase proportionally to $\sqrt{f}$ due to the skin effect (see Fig. 11.5).
  Note: The ohmic loss is a function of the frequency f [Hz] and of the geometry of the conductor (diameter of a conductor and surface area).
- **Dielectric Loss.** *Dielectric losses* occur because dielectric materials absorb energy from the propagating electric field (heating the material; see Sect. 7.8). The dielectric loss is a function of frequency f [Hz], the dissipation factor $\tan(\delta)$, and the dielectric constant ϵ_r'.
  Note: The dielectric loss does not depend on the geometry of the transmission line, only on the dielectric material.

7.8 Loss Tangent tan(δ)

The electric and magnetic losses quantify a material's inherent dissipation of electromagnetic energy (e.g., heat). The *loss tangent* tan(δ) (Fig. 7.15) is used to express how lossy a medium or a transmission line is: a small tan(δ) means low loss.

The electric loss tangent—often called dielectric loss tangent or *dissipation factor* D_f—is defined as [2]:

$$\tan(\delta_e) = \frac{\epsilon''}{\epsilon'} = \frac{\text{loss current}}{\text{charging current}} \tag{7.69}$$

where,

$\epsilon' = \epsilon'_r \epsilon_0$ = ability to store energy in a medium when an external electric field is applied = real part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]

$\epsilon'' = \epsilon''_r \epsilon_0$ = dielectric loss = energy dissipated when a medium is under the influence of an external electric field = imaginary part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]

The electrical conductivity σ [S/m] of a dielectric material is defined as [2]:

$$\sigma = \omega\epsilon'' \tag{7.70}$$

and therefore the loss tangent can be written as function of σ [S/m]:

$$\tan(\delta_e) = \frac{\epsilon''}{\epsilon'} = \frac{\sigma}{\omega\epsilon'} \tag{7.71}$$

The magnetic loss tangent tan(δ_m) is usually not of interest for traditional transmission lines because the materials involved are not magnetic. However, for the sake of completeness, here the definition of the magnetic loss tangent [2]:

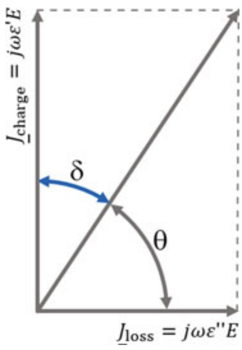


Fig. 7.15 Charging and loss current density

$$\tan(\delta_m) = \frac{\mu''}{\mu'} \tag{7.72}$$

where,

$\mu' = \mu'_r \mu_0$ = ability to store energy in a medium when an external magnetic field is applied = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]
$\mu'' = \mu''_r \mu_0$ = magnetic loss = the energy dissipated when a medium is under the influence of an external magnetic field = imaginary part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [F/m]

7.9 Balanced vs. Unbalanced Transmission Lines

EMC design engineers should be well aware of the concept and benefits of *balanced transmission lines*. Here the differences between *balanced* and *unbalanced transmission lines*:

- **Balanced.** A balanced transmission line consists of two conductors which have the same impedance along their line and the same impedance to ground and all other conductors (Fig. 7.16). Differential signals should be transmitted over balanced transmission lines.
- **Unbalanced.** In an unbalanced transmission line, the impedances of the forward and return current lines to ground are unequal. Single-ended signals are signals which are referenced to ground and should therefore be sent over unbalanced transmission lines (Fig. 7.17).

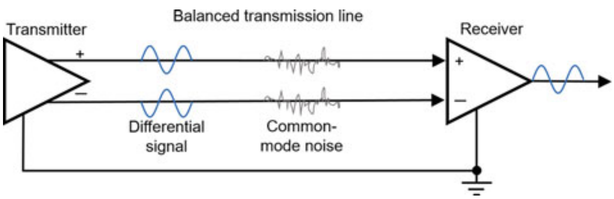


Fig. 7.16 Balanced transmission line with common-mode noise

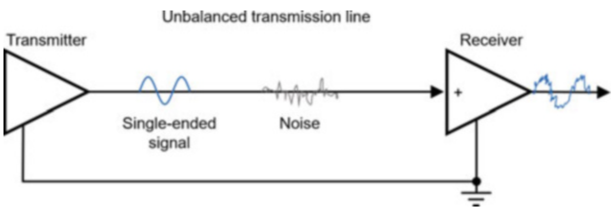


Fig. 7.17 Unbalanced transmission line with noise on signal line

Balanced transmission lines are very robust against common-mode noise because common-mode signals will be canceled out at the receiver's side. Common-mode noise coupling is explained in Sect. 12.4. However, to prevent differential-mode noise coupling, the two signal conductors must be routed close to each other (e.g., twisting them). Differential-mode noise coupling is explained in Sect. 12.3.

Generally, *single-ended interfaces* should be sent over *unbalanced transmission lines*, and *differential signal interfaces* should be sent over *balanced transmission lines*. Examples:

- **Unbalanced.** Suitable for single-ended signal interfaces.
  - **PCB data lines.** Microstrip lines (see Fig. 7.6), striplines (see Fig. 7.8), and coplanar waveguides (see Fig. 7.7)
  - **Cables and wires.** Coaxial cables (see Fig. 7.5) and multi-layer flat-ribbon or flat-flex cables with at least one solid ground plane
- **Balanced.** Suitable for differential and pseudo-differential signal interfaces.
  - **PCB data lines.** Microstrip lines and striplines routed as differential pairs
  - **Cables and wires.** Twisted pair (shielded or unshielded, see Fig. 7.4), twin-lead cables, flat-ribbon or flat-flex cables (one layer)

7.10 Single-Ended vs. Differential Interfaces

Single-ended signal interfaces consist of a single signal line and a reference potential. In contrast, *differential signal interfaces* consist of two complementary signals—a differential pair—and a reference potential (compare Figs. 7.18 and 7.19).

While Ethernet is a truly differential signal interface, most of the differential interfaces listed below are pseudo-differential, meaning that two single-ended digital signal sources are switched inverted relative to each other (around a common mode voltage level). The advantage of a pseudo-differential interface to a truly differential is that it doesn't need a *balun* (transformer which converts between a balanced signal and an unbalanced signal; see Sect. 11.7). The disadvantage of a pseudo-differential interface over a truly differential interface is that due to

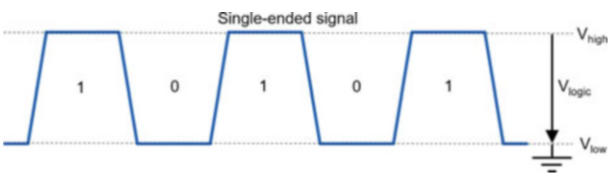


Fig. 7.18 Single-ended signal

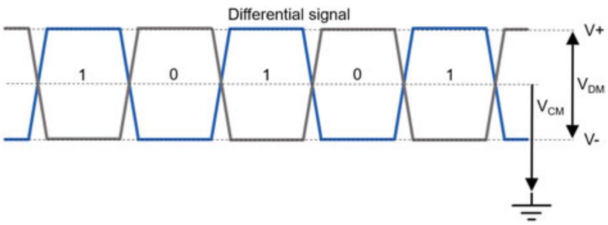


Fig. 7.19 Differential signal, where V_{DM} [V] is the differential signal voltage and V_{CM} [V] the common-mode voltage

slight timing differences at signal changes (from high to low and vice versa), a common-mode noise current is generated, which could lead to unintended radiated emissions (common-mode noise is the number one source for unintended radiation; see Sect. 9.9).

- **Single-ended interfaces.** Single-ended signals should be sent over unbalanced transmission lines.
  - **Communication interfaces.** RS232, VGA video connectors, SCSI interfaces, VMEbus, PCI
  - **PCB data buses.** I^2C, SPI
  - **Digital signaling.** TTL, CMOS
- **Differential signal interfaces.** Differential signals should be sent over balanced transmission lines.
  - **Communication interfaces.** Ethernet, USB, HDMI, DVI-D, Serial ATA, PCI Express
  - **Digital signaling.** LVDS, ECL, PECL, CML/SCL

7.11 Summary

- **Transmission line.** A transmission line is a series of conductors, often but not necessarily two, used to guide electromagnetic energy from one place to the other. Transmission line parameters are summarized in Table 7.7.
- **Critical length $\mathbf{l_{critical}}$.** Reflections and ringing can occur if a signal interconnection is longer than the critical length, and impedance matching must be considered to avoid this. Good and poor impedance matching is presented in Table 6.2. Rule of thumb:

$$l_{\text{critical}} \approx \frac{\lambda_{min}}{10} \tag{7.73}$$

Table 7.7 Transmission line parameters for any media, lossless media ($\sigma = 0$), low-loss media good ($\epsilon''/\epsilon' = \sigma/(\omega\epsilon') << 0.01$)) and good conductors $\epsilon''/\epsilon' > 100$) [5]

Transmission Line Parameters:
Propagation Constant $\gamma = \alpha + j\beta$. Attenuation Constant α. Phase Constant β. Intrinsic Impedance η. Wavelength λ. Phase Velocity v_p.

Parameter	Any media	Lossless medium ($\sigma = 0$)	Low-loss medium ($\varepsilon''/\varepsilon' << 1$)	Good conductor ($\varepsilon''/\varepsilon' >> 1$)
α [1/m]	$\omega\sqrt{\frac{\varepsilon'\mu' - \varepsilon''\mu''}{2}\left(1 + \sqrt{1 + \left(\frac{\varepsilon'\mu'' + \varepsilon''\mu'}{\varepsilon'\mu' - \varepsilon''\mu''}\right)^2} - 1\right)}$	0	$\frac{\sigma}{2}\sqrt{\frac{\mu'}{\varepsilon'}}$	$\sqrt{\pi f \mu' \sigma}$
β [rad/m]	$\frac{2\pi}{\lambda} = \omega\sqrt{\frac{\varepsilon'\mu' - \varepsilon''\mu''}{2}\left(\sqrt{1 + \left(\frac{\varepsilon'\mu'' + \varepsilon''\mu'}{\varepsilon'\mu' - \varepsilon''\mu''}\right)^2} + 1\right)}$	$\omega\sqrt{\mu'\varepsilon'}$	$\omega\sqrt{\mu'\varepsilon'}$	$\sqrt{\pi f \mu' \sigma}$
η [Ω]	$\sqrt{\frac{\mu''\omega + j\omega\mu'}{\varepsilon''\omega + j\omega\varepsilon'}}$	$\sqrt{\frac{\mu'}{\varepsilon'}}$	$\sqrt{\frac{\mu'}{\varepsilon'}}$	$(1 + j)\sqrt{\frac{\pi f \mu'}{\sigma}}$
λ [m]	$\frac{2\pi}{\beta} = \frac{v}{f} = \frac{1}{f\sqrt{\frac{\varepsilon'\mu' - \varepsilon''\mu''}{2}\left(\sqrt{1 + \left(\frac{\varepsilon'\mu'' + \varepsilon''\mu'}{\varepsilon'\mu' - \varepsilon''\mu''}\right)^2} + 1\right)}}$	$\frac{1}{f\sqrt{\mu'\varepsilon'}}$	$\frac{1}{f\sqrt{\mu'\varepsilon'}}$	$\sqrt{\frac{4\pi}{f\mu'\sigma}}$
v_p [m/s]	$\frac{\omega}{\beta} = \lambda f = \frac{1}{\sqrt{\frac{\varepsilon'\mu' - \varepsilon''\mu''}{2}\left(\sqrt{1 + \left(\frac{\varepsilon'\mu'' + \varepsilon''\mu'}{\varepsilon'\mu' - \varepsilon''\mu''}\right)^2} + 1\right)}}$	$\frac{1}{\sqrt{\mu'\varepsilon'}}$	$\frac{1}{\sqrt{\mu'\varepsilon'}}$	$\sqrt{\frac{4\pi f}{\mu'\sigma}}$

Notes:
$\varepsilon' = \varepsilon_r\varepsilon_0$. $\varepsilon'' = \sigma/\omega$. $\omega = 2\pi f$. Low-loss medium in practice: $\varepsilon''/\varepsilon' = \sigma/(\omega\varepsilon) < 0.01$. Good conducting medium in practice: $\varepsilon''/\varepsilon' > 100$

where,

l_{critical} = critical length in [m]
λ_{min} = wavelength of the highest significant harmonic frequency in the signal in [m]

- **Characteristic impedance $\underline{Z}_0$.** The characteristic impedance $\underline{Z}_0$ [Ω] of a uniform transmission line is the ratio of the complex amplitudes of voltage $\underline{V}$ [V] and current $\underline{I}$ [A] of a single wave propagating along the line in the same direction (in the absence of reflections in the other direction).

$$\underline{Z}_0 = \sqrt{\frac{R' + j\omega L'}{G' + j\omega C'}} \quad (7.74)$$

where,

R' = resistance per-unit-length in [Ω/m]
L' = inductance per-unit-length in [H/m]
C' = capacitance per-unit-length in [F/m]
G' = conductance per-unit-length in [S/m]
$\omega = 2\pi f$ = angular frequency in [rad/sec]
$j = \sqrt{-1}$ = imaginary unit

For a lossless line ($R' = 0$, $G' = 0$), the characteristic impedance is reduced to:

$$Z_0 = \sqrt{\frac{L'}{C'}} \tag{7.75}$$

- Rule of thumb for the **per-unit-length loop self-inductance L′** and **capacitance C′** for small signal cables (AWG6 to AWG32) and narrow signal lines on PCBs (0.1 mm to 0.25 mm):
  - $L' = 1\,\mu\text{H/m} = 1\,\text{nH/mm}$
  - $C' = 50\,\text{pF/m} = 50\,\text{fF/mm}$
- **Propagation constant $\underline{\gamma}$.** The propagation constant $\underline{\gamma}$ [1/m] of a sinusoidal electromagnetic wave is a measure of the change undergone by the amplitude and phase of the wave as it propagates along a transmission line in direction z [m].

$$\underline{\gamma} = \alpha + j\beta = \sqrt{(R' + j\omega L')(G' + j\omega C')} = j\omega\sqrt{\underline{\epsilon}\underline{\mu}} \tag{7.76}$$

where,

α = attenuation constant (or attenuation factor) in [1/m]
β = phase constant (or phase factor) in [rad/m]
$j = \sqrt{-1}$ = imaginary unit
R' = resistance per-unit-length in [Ω/m]
L' = inductance per-unit-length in [H/m]
C' = capacitance per-unit-length in [F/m]
G' = conductance per-unit-length in [S/m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\underline{\mu} = \mu' - j\mu''$ = complex permeability of the media through which the electromagnetic wave is propagating in [H/m]
$\underline{\epsilon} = \epsilon' - j\epsilon''$ = complex permittivity of the media through which the electromagnetic wave is propagating in [F/m]

- **Balanced vs. unbalanced.** Balanced transmission lines are more robust against EMI (common-mode interference) than unbalanced transmission lines.
- **Single-ended vs. differential.** Differential signals should be sent over balanced transmission lines and single-ended signals over unbalanced transmission lines.

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Chapter 8
Electromagnetic Fields

Thoroughly conscious ignorance is the prelude to every real advance in science.

—James Clerk Maxwell

8.1 The Electromagnetic Field

The *electromagnetic field* can be seen as the combination of an electric field $\overrightarrow{E}$ [V/m] and a magnetic field $\overrightarrow{H}$ [A/m] traveling in the same direction. Figure 8.1 shows an electromagnetic field in form of a plane wave, where the electric field and the magnetic field are perpendicular to each other.

The *Poynting vector*[1] $\overrightarrow{S}$ [W/m^2] represents the direction of propagation of an electromagnetic wave and the complex power density vector of a radiated electromagnetic field in [W/m^2]. $\overrightarrow{S}$ is defined as the cross product of the two vector fields $\overrightarrow{E}$ and $\overrightarrow{H}^*$ [1]:

$$\overrightarrow{S} = \overrightarrow{E} \times \overrightarrow{H}^* \tag{8.1}$$

where,

$\overrightarrow{S}$ = the directional energy flux vector (the energy transfer per unit area per unit time) of an electromagnetic field in [W/m^2]
$\overrightarrow{E}$ = electric field vector in [V/m]
$\overrightarrow{H}^*$ = complex conjugate of the magnetic field vector $\overrightarrow{H}$ in [A/m]

Instead of working with complex field vectors, it is often much more practical to work with the average power density S_{avg} [W/m^2]. For a uniform plane wave

[1] The Poynting vector represents the directional energy flux (the energy transfer per unit area per unit time) of an electromagnetic field. The SI unit of the Poynting vector is the watt per square meter [W/m^2]. It is named after its discoverer John Henry Poynting who first derived it in 1884.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_8

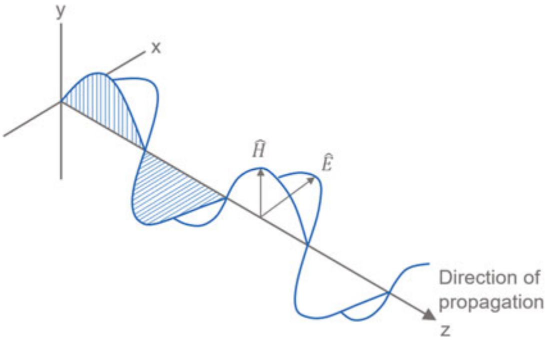

Fig. 8.1 An electromagnetic field (plane wave) can be seen as the combination of an electric field and a magnetic field (perpendicular to each other and traveling in the same direction)

(electromagnetic field in the far-field, where the wave impedance $\underline{Z}_w$ [Ω] is equal to the intrinsic impedance $\underline{\eta}$) in a lossless medium ($\underline{\epsilon}_r = 1.0$ and $\underline{\mu}_r = 1.0$), the calculation of the average power density S_{avg} [W/m^2] can be simplified to:

$$S_{avg} = \frac{1}{2}|\vec{E}| \cdot |\vec{H}| = \frac{|\vec{E}|^2}{2\eta} = \frac{|\vec{H}|^2\eta}{2} \tag{8.2}$$

where,

$|\vec{E}|$ = amplitude (magnitude) of the electric field vector/phasor in [V/m]
$|\vec{H}|$ = amplitude (magnitude) of the magnetic field vector/phasor in [A/m]
η = intrinsic impedance of the medium where the uniform plane wave is traveling though in [Ω]

In Eq. 8.2 the factor $1/2$ can be omitted for RMS values of $\vec{E}$ [V/m] and $\vec{H}$ [A/m]. For free space, η [Ω] can be set to $\eta_0 = 377\,\Omega$. More details about the electromagnetic and other physical fields can be found in Appendix D.

8.2 Electromagnetic Field Characteristics

Three parameters determine the electromagnetic field characteristics [4]:

- **Source.** The physics of an antenna determines the wave impedance $\underline{Z}_w$ [Ω] and the distance of the near-field/far-field boundary from the antenna. An antenna can be an intentional antenna like a dipole, horn or loop antenna, or an unintended antenna like a cable or a PCB trace.
- **Media.** Every medium has its intrinsic impedance $\underline{\eta}$ [Ω]. The medium that surrounds the source (e.g., air, plastics, metal) and the medium through which the electromagnetic wave is traveling influence the wave impedance $\underline{Z}_w$ [Ω] and the attenuation.
- **Distance.** The distance between the source and the point of observation is an essential factor. Close (compared to the wavelength λ [m]) to the source, the field

properties are determined primarily by the source characteristics (low-impedance or high-impedance source). Far from the source, the field depends mainly on the medium through which the field is propagating. Therefore, the space around a radiation source (antenna) can be split into two regions: the near-field and the far-field.

More details about the near-field, the far-field, the wave impedance $\underline{Z}_w$ [Ω], and the intrinsic impedance $\underline{\eta}$ [Ω] are presented in the following Sects. 8.3–8.5.

8.3 Near-Field vs. Far-Field

As an EMC design engineer, there is no way around the near- and far-field topic. For example, for shielding, it is crucial whether the electromagnetic wave to be shielded is a near-field or a far-field wave (see Chap. 13). Moreover, for EMC emission troubleshooting, it is crucial to know if a measurement takes place in the near-field or the far-field because different probes and antennas have to be used accordingly:

- **Measurement in the near-field.** Special near-field probes are used (see Fig. 9.3 on page 112).
- **Measurement in the far-field.** Typically, E-field antennas are used (see Fig. 9.1 on page 112).

The electromagnetic field around an antenna can be divided into three regions [1]:

1. Reactive near-field (see Sect. 8.3.1.1)
2. Radiating near-field (see Sect. 8.3.1.2)
3. Far-field (see Sect. 8.3.2)

The regions depend on the maximum linear dimension of the antenna D [m] and the wavelength of the signal λ [m] (see Fig. 8.2):

- **Electrically small antennas: D $<$ $\lambda/(2\pi)$.** The reactive near-field is significant. For electrically small antennas, the radiating near-field and the far-field are minimal, if they exist at all.
- **Electrically large antennas: D $>$ $\lambda/(2\pi)$.** All three regions are significant: the reactive near-field, the radiating near-field, and the far-field.

8.3.1 Near-Field

In the *near-field*—also called *Fresnel zone*—the wave impedance $\underline{Z}_w$ [Ω] depends primarily on the source, and the electric and magnetic fields have to be considered separately (because the ratio of $|\underline{E}(z)|/|\underline{H}(z)|$ is not constant). Usually, there is

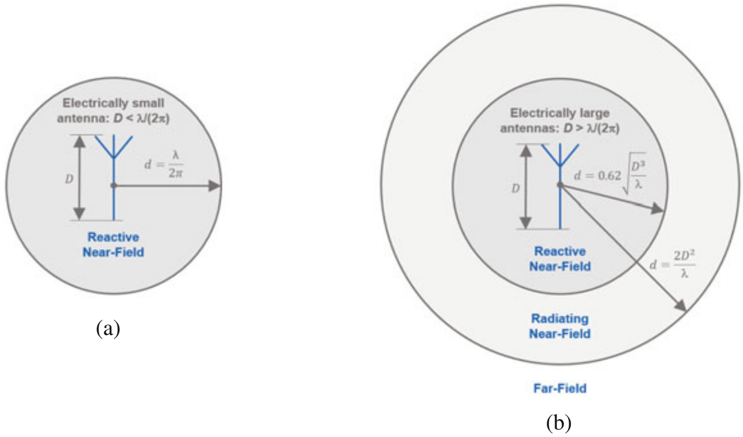

Fig. 8.2 Approximate near-field to far-field boundary [1]. (**a**) The reactive near-field of electrically small antennas. (**b**) The near-/ and far-field of electrically large antennas

either the electric E-field or the magnetic H-field predominant in the near-field [5]:

- **Predominant E-field:**
  - The source voltage is high compared to the source current (electric dipole).
  - The source impedance is high (e.g., electric dipole antennas).
  - The wave impedance near the antenna is high.
  - E-field attenuates with a rate of $1/d^3$ in the near-field (d = distance to source).
  - H-field attenuates with a rate of $1/d^2$ in the near-field (d = distance to source).
- **Predominant H-field:**
  - The source voltage is low compared to the source current.
  - The source impedance is low (e.g., loop antennas, magnetic dipole antennas).
  - The wave impedance near the antenna is low.
  - E-field attenuates with a rate of $1/d^2$ in the near-field (d = distance to source).
  - H-field attenuates with a rate of $1/d^3$ in the near-field (d = distance to source).

As mentioned above, the near-field can be divided into the following two regions: the reactive near-field and the radiating near-field.

8.3.1.1 Reactive Near-Field

In the *reactive near-field*, energy is stored in the electric and magnetic fields very close to the source but not radiated from them. Instead, energy is exchanged between the signal source and the fields.

In the case of $D < \lambda/(2\pi)$—this means that the antenna is not a very effective radiator—the reactive near-field extends until the distance d [m] from the antenna by Balanis [1]:

$$d_{reactive-near-field} \leq \frac{\lambda}{2\pi} \tag{8.3}$$

In the case of $D > \lambda/(2\pi)$—where there is the chance that the antenna is an effective radiator—the reactive near-field extends until the distance d [m] from the antenna by:

$$d_{reactive-near-field} \leq 0.62\sqrt{\frac{D^3}{\lambda}} \tag{8.4}$$

where,

λ = wavelength of the sinusoidal signal in [m]
D = maximum linear dimension of the antenna in [m]. $D = l$ for a wire antenna.

8.3.1.2 Radiating Near-Field

If a *radiating near-field* exists, then it is defined as the region between the reactive near-field and the far-field. In the radiative or radiating near-field, the angular field distribution depends on distance d from the antenna, unlike in the far-field where it does not depend on the distance. In addition, in the radiating near-field, the radiating power density is greater than the reactive power density.

If the antenna has a maximum dimension D [m] that is small compared to the wavelength λ [m], the radiating near-field region may not exist.

In the case of $D > \lambda/(2\pi)$, the radiating near-field begins after the reactive near-field region has ended ($d_{reactive-near-field}$) and ends where the far-field begins [1]:

$$d_{radiating-near-field} = 0.62\sqrt{\frac{D^3}{\lambda}} \ldots \frac{2D^2}{\lambda} \tag{8.5}$$

where,

λ = wavelength of the sinusoidal signal in [m].
D = maximum linear dimension of the antenna in [m]. $D = l$ for a wire antenna.

8.3.2 Far-Field

In the *far-field*—also called *Fraunhofer region*—the E- and H-fields move perpendicular (orthogonal) and in phase to each other and form a plane wave.

- E- and H-field attenuate with a rate of $1/d$ in the far-field (d = distance to source) and therefore the power density S_{avg} [W/m^2] of the electromagnetic wave attenuates with $1/d^2$.
- The wave impedance in free-space (air) is $\eta_0 = 377\,\Omega$.

There is usually only a far-field region in the case of $D > \lambda/(2\pi)$, where λ [m] is the wavelength of the signal and D [m] is the maximum linear dimension of the antenna. However, if there is a far-field, it starts roughly at the following distance d [m] from the antenna [1]:

$$d_{far-field} > \frac{2D^2}{\lambda} \tag{8.6}$$

where,

λ = wavelength of the sinusoidal signal in [m]
D = maximum linear dimension of the antenna in [m]. $D = l$ for a wire antenna.

8.4 Intrinsic Impedance η

8.4.1 Intrinsic Impedance of Any Media

The *intrinsic impedance* $\underline{\eta}$ [Ω] is a property of a medium, and it influences the electromagnetic waves that are traveling through that medium. The intrinsic impedance is dependent on the conductivity, permittivity and permeability of the medium. It is a complex number and defined as [6]:

$$\underline{\eta} = \sqrt{\frac{\underline{\mu}}{\underline{\epsilon}}} = \sqrt{\frac{\mu' - j\mu''}{\epsilon' - j\epsilon''}} = \sqrt{\frac{\mu'' + j\mu'}{\epsilon'' + j\epsilon'}} = \sqrt{\frac{\mu''\omega + j\omega\mu'}{\epsilon''\omega + j\omega\epsilon'}} \tag{8.7}$$

where,

$\underline{\mu} = \underline{\mu}_r \mu_0 = \mu' - j\mu''$ = complex permeability of the medium in [H/m]
$\underline{\epsilon} = \underline{\epsilon}_r \epsilon_0 = \epsilon' - j\epsilon''$ = complex permittivity of the medium in [F/m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$j = \sqrt{-1}$ = imaginary unit

8.4.2 Intrinsic Impedance of Magnetic Lossless Media

If we neglected the magnetic losses ($\mu'' = 0$) and set the dielectric loss to $\epsilon'' = \sigma/\omega$, we can write the intrinsic impedance of a medium like this [2]:

$$\underline{\eta} = \sqrt{\frac{j\omega\mu'}{\sigma + j\omega\epsilon'}} \tag{8.8}$$

where,

$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\mu' = \mu_r' \mu_0$ = permeability of the magnetic lossless medium in [H/m]
σ = specific conductance of the magnetic lossless medium in [S/m]
$\epsilon' = \epsilon_r' \epsilon_0$ = permittivity (dielectric constant) of the dielectric lossless medium in [F/m]
$j = \sqrt{-1}$ = imaginary unit

8.4.3 Intrinsic Impedance of Lossless Insulators

The calculation of the intrinsic impedance η [Ω] can even be more simplified for good insulators and lossless media ($\sigma \ll j\omega\epsilon'$) [2]:

$$\eta = \sqrt{\frac{\mu'}{\epsilon'}} \tag{8.9}$$

where,

$\mu' = \mu_r' \mu_0$ = permeability of the lossless media in [H/m]
$\epsilon' = \epsilon_r' \epsilon_0$ = permittivity (dielectric constant) of the lossless media in [F/m]

8.4.4 Intrinsic Impedance of Free-Space

For free space—where $\mu_r' = 1.0$ and $\epsilon_r' = 1.0$—the intrinsic impedance is [6]:

$$\eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \approx 377\,\Omega \tag{8.10}$$

where,

$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity

8.4.5 Intrinsic Impedance of Good Conductors

For waves traveling through (not along: through, e.g., through a shield) good conductors—where $\sigma \gg \epsilon_0 \omega$—the intrinsic impedance is [6]:

$$\underline{\eta} = (1+j)\sqrt{\frac{\omega\mu'}{2\sigma}} = (1+j)\frac{1}{\delta\sigma} \tag{8.11}$$

where,

$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\mu' = \mu_r' \mu_0$ = permeability of the conductive material in [H/m]
σ = specific conductance of the conductive material in [S/m]
$\delta = 1/\sqrt{\pi f \mu_r' \mu_0 \sigma}$ = skin depth in [m]

8.5 Wave Impedance Z_w

8.5.1 Wave Impedance of Any Wave

Generally, the ratio of the E-field [V/m] to the H-field [A/m]—for any electromagnetic wave—is the *wave impedance* $\underline{Z}_w$ [Ω]. Because of the vectorial character of the fields, the ratio must be defined in terms of the corresponding x-component $\underline{Z}_{wx}$ [Ω] and y-component $\underline{Z}_{wy}$ [Ω] of the $\underline{E}(z)$ [V/m] and $\underline{H}(z)$ [A/m] fields [3]:

$$\underline{Z}_{wx}(z) = \frac{\left[\vec{E}(z)\right]_x}{\left[\vec{H}(z) \times \vec{z}\right]_x} = \frac{\underline{E}_x(z)}{\underline{H}_y(z)} \tag{8.12}$$

$$\underline{Z}_{wy}(z) = \frac{\left[\vec{E}(z)\right]_y}{\left[\vec{H}(z) \times \vec{z}\right]_y} = -\frac{\underline{E}_y(z)}{\underline{H}_x(z)} \tag{8.13}$$

where,

$\vec{E}(z)$ = electric field vector traveling in z-direction in [V]
$\left[\vec{E}(z)\right]_x = \underline{E}_x(z)$ = electric field x-component of the wave traveling in z direction as complex phasor in [V]
$\left[\vec{E}(z)\right]_y = \underline{E}_y(z)$ = electric field y-component of the wave traveling in z direction as complex phasor in [V]
$\vec{H}(z)$ = magnetic field vector traveling in z-direction in [A/m]

$\left[\overrightarrow{H}(z) \times \overrightarrow{z}\right]_x = \underline{H}_y(z) =$ magnetic field y-component of the wave traveling in z direction as complex phasor in [A/m]
$\left[\overrightarrow{H}(z) \times \overrightarrow{z}\right]_y = \underline{H}_x(z) =$ magnetic field x-component of the wave traveling in z direction as complex phasor in [A/m]
$\overrightarrow{z} =$ unity vector in z-direction

The wave impedance of the x-component $\underline{Z}_{wx}(z)$ and y-component $\underline{Z}_{wy}(z)$ can also be written with respect to the intrinsic impedance of the medium $\underline{\eta}$ and the amplitudes E_{0x+} of the forward wave of the electric field x-component, E_{0y+} of the forward wave of the electric field y-component, E_{0x-} of the backward wave of the electric field x-component, and E_{0y-} of the backward wave of the electric field y-component [3]:

$$\underline{Z}_{wx}(z) = \frac{\underline{E}_x(z)}{\underline{H}_y(z)} = \underline{\eta}\frac{E_{0x+}e^{-\underline{\gamma}z} + E_{0x-}e^{\underline{\gamma}z}}{E_{0x+}e^{-\underline{\gamma}z} - E_{0x-}e^{\underline{\gamma}z}} \tag{8.14}$$

$$\underline{Z}_{wy}(z) = -\frac{\underline{E}_y(z)}{\underline{H}_x(z)} = -\underline{\eta}\frac{E_{0y+}e^{-\underline{\gamma}z} + E_{0y-}e^{\underline{\gamma}z}}{E_{0y+}e^{-\underline{\gamma}z} - E_{0y-}e^{\underline{\gamma}z}} \tag{8.15}$$

where,

$\underline{\gamma} = j\omega\sqrt{\underline{\mu\epsilon}} =$ complex propagation constant in [1/m]
$\underline{E}_x(z) = E_{0x+}e^{-\underline{\gamma}z} + E_{0x-}e^{\underline{\gamma}z} =$ electric field x-component, propagating in z direction in [V/m]
$\underline{E}_y(z) = E_{0y+}e^{-\underline{\gamma}z} + E_{0y-}e^{\underline{\gamma}z} =$ electric field y-component, propagating in z direction in [V/m]
$\underline{H}_x(z) = -\frac{1}{\underline{\eta}}\left[E_{0y+}e^{-\underline{\gamma}z} - E_{0y-}e^{\underline{\gamma}z}\right] =$ magnetic field x-component, propagating in z direction in [A/m]
$\underline{H}_y(z) = \frac{1}{\underline{\eta}}\left[E_{0x+}e^{-\underline{\gamma}z} - E_{0x-}e^{\underline{\gamma}z}\right] =$ magnetic field y-component, propagating in z direction in [A/m]

8.5.2 Wave Impedance vs. Distance

Figure 8.3 shows the wave impedance $|\underline{Z}_w|$ [Ω] in dependency of the distance d [m] from the radiation source (normalized to the near-/far-field boundary). The graph is a simplification, and it should illustrate that magnetic field antennas and electric field antennas have different wave impedances in the near-field and that the electromagnetic field in the far-field has a constant wave impedance (for free space: $Z_w = \eta_0 \approx 377\,\Omega$).

The formulas in Sect. 8.5 showed that the general calculation of the wave impedance $\underline{Z}_w$ is complicated. However, this changes if we solely focus on the

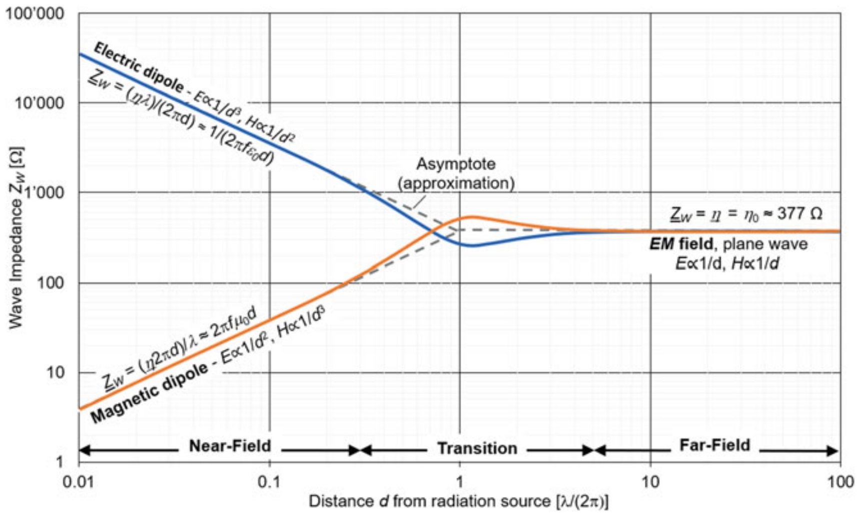

Fig. 8.3 Wave impedance $|\underline{Z}_w|$ vs. distance d from source (antenna)

far-field: in the far-field, the wave impedance is equal to the intrinsic impedance of the material through which the wave is propagating ($\underline{Z}_w = \underline{\eta}$).

In the near-field, the wave impedance depends primarily on the source. Electromagnetic waves are generated by two types of sources:

- **High-impedance sources, E-field antennas.** For high-impedance sources, the wave impedance $|\underline{Z}_w|$ [Ω] in the near-field is high and the E-field dominates. Examples of high-impedance sources are all kinds of wireless communication antennas (5G, Bluetooth, WiFi, RFID), radar antennas, and unintended antennas like cables, wires, and PCB traces. The frequency range of high-impedance antennas is very broad: from 10 kHz up to 100 GHz.
- **Low-impedance sources, H-field antennas.** For low-impedance sources, the wave impedance $|\underline{Z}_w|$ [Ω] in the near-field is low, and the H-field dominates. Examples of low-impedance sources are transformers, motors, inductors, coils, inductive charging systems, current loops, and all kinds of conductors with high currents. The frequency range of low-impedance antennas is typically not very broad: from 50 Hz (transformers) to several 100 kHz (inductive charging) up to 10 MHz (DC/DC converters).

8.5.3 Wave Impedance in the Near-Field of E-Field Antennas

The wave impedance $\underline{Z}_{we}$ [Ω] in the near-field of E-field antennas is approximately [5]:

$$\underline{Z}_{we} \approx -j\frac{\underline{\eta}}{\beta d} = -j\frac{\underline{\eta}\lambda}{2\pi d} \tag{8.16}$$

where,

$\underline{Z}_{we}$ = wave impedance in the near-field of an E-field antenna in [Ω]
$\underline{\eta}$ = intrinsic impedance of media where wave is propagating through in [Ω]
$\beta = 2\pi/\lambda$ = phase constant (or phase factor) of the sinusoidal wave in [rad/m]
λ = wavelength of the sinusoidal wave in [m]
d = distance from the E-field antenna in [m]

In case of free space, where $\underline{\eta} = \eta_0 = \sqrt{\mu_0/\epsilon_0}$ and $\lambda = 1/(\sqrt{\mu_0\epsilon_0} f)$, the wave impedance in the near-field of E-field antennas can be simplified to [4]:

$$|Z_{we}| \approx \frac{1}{2\pi f \epsilon_r' \epsilon_0 d} = \frac{1}{2\pi f \epsilon_0 d} \tag{8.17}$$

where,

$|Z_{we}|$ = wave impedance in the near-field of a E-field antenna in [Ω]
f = frequency of the sinusoidal signal in [Hz]
ϵ_r' = relative electric permittivity of the medium in the near-field
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity
d = distance from the E-field antenna in [m]

8.5.4 Wave Impedance in the Near-Field of H-Field Antennas

The wave impedance $\underline{Z}_{wm}$ [Ω] in the near-field of H-field antennas is approximately [5]:

$$\underline{Z}_{wm} \approx -j\underline{\eta}\beta d = -j\frac{\underline{\eta}2\pi d}{\lambda} \tag{8.18}$$

where,

$\underline{Z}_{wm}$ = wave impedance in the near-field of a H-field antenna in [Ω]
$\underline{\eta}$ = intrinsic impedance of media where wave is propagating through in [Ω]
$\beta = 2\pi/\lambda$ = phase constant (or phase factor) of the sinusoidal wave in [rad/m]
λ = wavelength of the sinusoidal wave in [m]
d = distance from the H-field antenna in [m]

In case of free space, where $\underline{\eta} = \eta_0 = \sqrt{\mu_0/\epsilon_0}$) and $\lambda = 1/(\sqrt{\mu_0\epsilon_0} f)$, the wave impedance in the near-field of H-field antennas can be simplified to [4]:

$$|Z_{wm}| \approx 2\pi f \mu_r' \mu_0 d = 2\pi f \mu_0 d \tag{8.19}$$

where,

$|Z_{wm}|$ = wave impedance in the near-field of a H-field antenna in [Ω]
f = frequency of the sinusoidal signal in [Hz]
μ_r' = relative magnetic permeability of the medium in the near-field
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
d = distance from the H-field antenna in [m]

8.5.5 Wave Impedance of Plane Waves

For plane waves—meaning electromagnetic waves in the far-field—the wave impedance $\underline{Z}_w$ [Ω] is equal to the intrinsic impedance $\underline{\eta}$ [Ω] of the medium:

$$\underline{Z}_w = \underline{\eta} \tag{8.20}$$

In case of free space, the wave impedance is simply:

$$Z_w = \eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \approx 377\,\Omega \tag{8.21}$$

where,

$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity

8.6 Summary

- **Electromagnetic field.** The electromagnetic field can be viewed as an interdependent time-varying field consisting of an electric field E and magnetic field H, propagating in the same direction.
- **Near- and far-field.** The area around an antenna can be divided into three regions: reactive near-field, radiative near-field, and far-field (Table 8.1).
  - **Electrically small antennas D $< \lambda/(2\pi)$.** Roughly speaking, only the reactive near-field is significant for electrically small antennas. It extends from the antenna to the distance $d_{reactive-near-field}$:

$$d_{reactive-near-field} = 0 \ldots \frac{\lambda}{2\pi} \tag{8.22}$$

where,

λ = wavelength of the sinusoidal signal in [m]

Table 8.1 Comparison of two criteria (Rayleigh and Maxwell) for the near-field to far-field transition for various frequencies f [Hz] and maximum antenna dimensions D [m]

Frequency	$\lambda/2$ [m]	Maximum dimension D [m]	Rayleigh $d = 2D^2/\lambda$ [m]	Maxwell $d = \lambda/(2\pi)$ [m]
30 MHz	5	10	20	1.59
30 MHz	5	2	0.8	1.59
100 MHz	1.5	3	6	0.48
100 MHz	1.5	0.5	0.17	0.48
300 MHz	0.5	1	2	0.16
300 MHz	0.5	0.1	0.02	0.16
1 GHz	0.15	0.3	0.6	0.05
1 GHz	0.15	0.05	0.017	0.05
3 GHz	0.05	0.1	0.2	0.02
3 GHz	0.05	0.02	0.008	0.02
6 GHz	0.025	0.05	0.1	0.01
6 GHz	0.025	0.01	0.004	0.01

– **Electrically large antennas D $> \lambda/(2\pi)$.** For electrically large antennas, all three regions are significant.

$$d_{reactive-near-field} = 0 \ldots 0.62\sqrt{\frac{D^3}{\lambda}} \tag{8.23}$$

$$d_{radiating-near-field} = 0.62\sqrt{\frac{D^3}{\lambda}} \ldots \frac{2D^2}{\lambda} \tag{8.24}$$

$$d_{far-field} > \frac{2D^2}{\lambda} \tag{8.25}$$

where,

λ = wavelength of the sinusoidal signal in [m]
D = maximum linear dimension of the antenna in [m]. $D = l$ for a wire antenna.

- **Intrinsic impedance $\underline{\eta}$.** The intrinsic impedance $\underline{\eta}$ depends only on the properties of the material (conductivity, permeability, permittivity), and for plane waves, the wave impedance $\underline{Z}_w$ is equal to the intrinsic impedance.

$$\underline{\eta} = \sqrt{\frac{\underline{\mu}}{\underline{\epsilon}}} \tag{8.26}$$

where,

$\underline{\mu} = \mu' - j\mu''$ = complex permeability of the media through which the electromagnetic wave is propagating in [H/m]
$\underline{\epsilon} = \epsilon' - j\epsilon''$ = complex permittivity of the media through which the electromagnetic wave is propagating in [F/m]

- **Wave impedance $\underline{\mathbf{Z}}_{\mathbf{w}}$.** The wave impedance is a characteristic of a particular wave (which depends on the source antenna, the material, the frequency, etc.). For plane waves, we get:

$$\underline{Z}_w = \underline{\eta} \tag{8.27}$$

where,

$\underline{\eta}$ = intrinsic impedance of the material through which the electromagnetic wave is propagating in [Ω]

And for plane waves in free space, we get:

$$Z_w = \eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} = 377\,\Omega \tag{8.28}$$

where,

$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
$\epsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity

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Chapter 9
Antennas

Antennas do not match transmission line impedances to wave impedances. If they did, a 377 Ω transmission line would radiate all its power without requiring an antenna. Describing an antenna as an impedance transformator is similar to telling a child that babies are delivered by storks. Both are explanations that may satisfy a certain audience, but neither is consistent with the actual physics.

— Dr. Todd Hubing

9.1 Antennas and EMC Testing

An *antenna* is an interface between radio waves propagating through space and electric currents moving along metal conductors. It can act as a receiver (converting electromagnetic waves from space into voltages and currents in a conductor) or as a transmitter (converting voltages and currents into electromagnetic waves).

In EMC compliance testing, antennas have two functions:

- **Radiated emission testing.** Antennas act as receivers.
- **Radiated immunity testing.** Antennas act as transmitters.

Moreover, a distinction is made between E-field and H-field antennas.

- **E-field antennas.** *Electric field antennas*, for example: biconical antennas, bilogical antennas, horn antennas, log-periodic antennas, and electric near-field probes (see Fig. 9.1).
- **H-field antennas.** *Magnetic field antennas* are usually all kinds of loop antennas for the near- and the far-field (see Fig. 9.2).

And last but not least, antennas are divided into near-field and far-field antennas:

- **Near-field antennas.** Dedicated *near-field antennas* are called near-field probes (see Fig. 9.3). Electric and magnetic near-field probes are often used during EMC troubleshooting to identify unintentional radiators on PCBAs or unintended

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_9

Fig. 9.1 E-field antenna examples by Rohde & Schwarz. From left to right: double-ridged waveguide horn antenna R&S®HF907, log-periodic antenna R&S®HL223, biconical antenna R&S®HK116E, and ultralog antenna R&S®HL562E

Fig. 9.2 H-field antenna examples by Rohde & Schwarz. From left to right: active loop antenna R&S®HFH2-Z2E, triple-loop antenna R&S®HM020E

Fig. 9.3 EMC near-field probes R&S®HZ-15 with oscilloscope R&S®RTO2000 by Rohde & Schwarz. H-field near-field probes have loops, E-field probes not

radiation through slots of enclosures. Magnetic field immunity tests are usually performed in the near-field of a loop antenna (e.g. at $f = 50\,\text{Hz}$ or $60\,\text{Hz}$).

- **Far-field antennas.** *Far-field antennas* are used for EMC compliance emission and immunity tests (see Fig. 9.1). In some cases, far-field antennas are also used for near-field radiated immunity tests (radiated fields in close proximity [6]).

9.2 Isotropic Radiator

A spherical *isotropic radiator* (isotropic antenna) radiates equally in all directions (from one single point) and therefore has no directivity. Such an antenna does only exist in theory.

Let us define S [W/m^2] as the power density of the field around an antenna at a given distance d [m]. The average *power density* S [W/m^2] of an isotropic radiator is simply the radiated power P_{rad} [W] by the antenna divided by the surface area of the spherical surface [m^2] at a distance d [m] from the center of the radiator [11]:

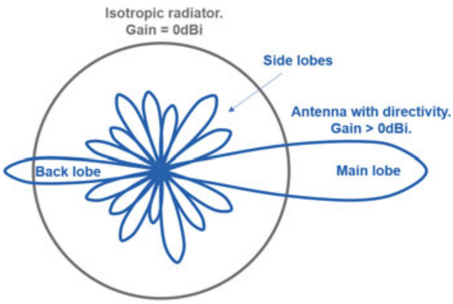


Fig. 9.4 Signal strength of a spherical isotropic radiator vs. an antenna with directivity (gain)

$$S = \frac{P_{rad}}{4\pi d^2} \tag{9.1}$$

where,

P_{rad} = radiated power by an isotropic radiator in [m^2]
d = distance from the isotropic radiator to the spherical surface in [m]

9.3 Antenna Directivity D

The *directivity D* of an antenna is defined as the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions. Thus, the directivity is an antenna parameter that measures the degree to which the emitted radiation is concentrated in a single direction. Figure 9.4 shows the radiation pattern (power density) of a transmitting antenna with directivity compared to an isotropic radiator. Directivity is also defined for receiving antennas. The directivity when receiving is equal to the directivity when transmitting.

9.4 Antenna Gain G

Every real-world antenna has a directivity—meaning: it does not radiate equally in every direction. This leads us to the term *antenna gain*. The antenna gain G is a key performance number that combines the antenna's directivity D and electrical efficiency e_0 [2]:

$$G = e_0 \cdot D \tag{9.2}$$

where,

e_0 = radiation efficiency of antenna (dimensionless)

D = directivity of antenna (without any losses, dimensionless)

For a loss-free antenna: $G = D$. The antenna gain G is defined as the ratio of the power radiated in the desired direction of an antenna compared to the power radiated from a reference antenna (e.g., isotropic radiator or dipole) with the identical power input (this means the antenna's electrical efficiency factor P_{rad}/P_t, which considers the antenna losses, is already taken into account in gain G) [11]. The antenna gain with reference to an isotropic radiator G_i or to a $\lambda/2$-dipole antenna G_d is defined as:

$$G_i = \frac{\text{Maximum radiation intensity of actual antenna}}{\text{Radiation intensity of isotropic antenna with same input power}} \tag{9.3}$$

$$G_d = \frac{\text{Maximum radiation intensity of actual antenna}}{\text{Radiation intensity of } \lambda/2\text{-dipole antenna with same input power}} \tag{9.4}$$

where,

G_i = gain of an antenna with reference to an isotropic radiator (dimensionless)
G_d = gain of an antenna with reference to a $\lambda/2$-dipole antenna (dimensionless)

Antenna gains are often given in [dBi] and [dBd]:

$$G_i[\text{dBi}] = 10\log_{10}(G_i) \tag{9.5}$$

$$G_d[\text{dBd}] = 10\log_{10}(G_d) \tag{9.6}$$

The antenna gains of loss-free $\lambda/2$-dipoles and $\lambda/4$-monopoles compared to an isotropic radiator are [2]:

$$G_{i,\lambda/2-dipole} = 1.64 \text{ loss-free antenna} \tag{9.7}$$

$$G_{i,\lambda/4-monopole} = 3.28 \text{ loss-free antenna} \tag{9.8}$$

$$G_{i,\lambda/2-dipole}[\text{dBi}], [\text{dBd}] = 10\log_{10}(1.64) = 2.15\,\text{dBi} = 0\,\text{dBd} \tag{9.9}$$

$$G_{i,\lambda/4-monopole}[\text{dBi}], [\text{dBd}] = 10\log_{10}(3.28) = 5.15\,\text{dBi} = 3\,\text{dBd} \tag{9.10}$$

Now, we are ready to calculate the often used terms: *effective isotropic radiated power* (EIRP) and effective radiated power (ERP) [9]:

$$\text{EIRP} = P_t G_i = 1.64 \cdot \text{ERP} \tag{9.11}$$

$$\text{ERP} = P_t G_d = \frac{\text{EIRP}}{1.64} \tag{9.12}$$

where,

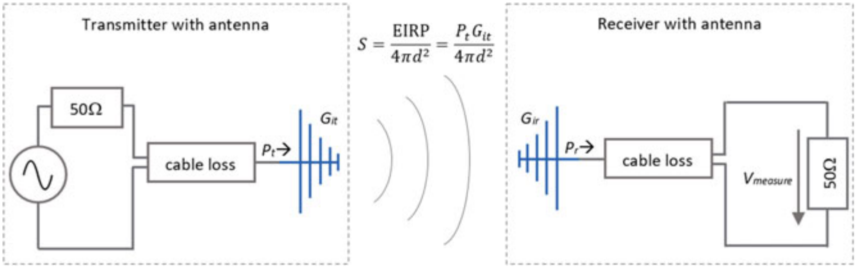


Fig. 9.5 Transmitting and receiving antennas in free-space

EIRP = RMS input power required to lossless isotropic radiator to give the identical maximum power density far from the antenna as the actual antenna in [W]
ERP = RMS input power required to lossless $\lambda/2$-dipole to give the identical maximum power density far from the antenna as the actual antenna in [W]
P_t = power input to the transmitting antenna in [W]
G_i = antenna gain compared to isotropic radiator (dimensionless)
G_d = antenna gain compared to $\lambda/2$-dipole (dimensionless)

Assuming line of sight and free space (Fig. 9.5), the maximum power density S_{max} [W/m^2] at a distance d [m] from the transmitting antenna (in the direction of the transmitting antenna's main-lobe and in the far-field) can be written as:

$$S_{max} = \frac{\text{EIRP}}{4\pi d^2} \tag{9.13}$$

where,

EIRP = RMS input power required to lossless isotropic radiator to give the identical maximum power density far from the antenna as the actual antenna in [W]
d = distance from the transmitting antenna (in the far-field) in [m]

The term *realized antenna gain* $G_{realized}$ does not only include the antenna efficiency (ohmic losses, etc.) but also the impedance mismatch loss and is given as [1]:

$$G_{realized} = G_i \cdot \left(1 - |\underline{\Gamma}|^2\right) \tag{9.14}$$

where,

G_i = antenna gain, compared to a spherical isotropic radiator (dimensionless)
$|\underline{\Gamma}|$ = magnitude of the complex reflection coefficient between the signal generator or amplifier and the antenna (dimensionless)

9.5 Effective Aperture A_e

The effective area of an antenna is also called effective aperture A_e [m^2]. The *effective aperture* represents the ratio of power P_r [W] (output power of the receiving antenna) to the power density S_r [W/m^2] of the electromagnetic wave at the receiving antenna [2]:

$$A_e = \frac{P_r}{S_r} \tag{9.15}$$

where,

P_r = RMS output power at the receiving antenna terminals in [W]
S_r = power density S_r at the receiving antenna in [W/m^2]

The *maximum effective aperture* A_{em} [m^2] for any antenna is [2]:

$$A_{em} = \frac{\lambda^2}{4\pi} G_i \tag{9.16}$$

where,

λ = wavelength of the sinusoidal incident wave—which must be matched with the receiving antenna's polarization—in [m]
G_i = antenna gain of the receiving antenna (dimensionless)

The antenna gain G_i can be expressed as a function of the maximum effective aperture A_{em} [m^2] and the wavelength λ [m] of the incident wave:

$$G_i = \frac{4\pi A_{em}}{\lambda^2} \tag{9.17}$$

9.6 Antenna Factor AF

The *antenna factor* AF is a function of the signal's frequency f [Hz] and is used to calculate the received field strength E [V/m] or H [A/m] based on the measured voltage V_r [V] at the receiving antenna's terminals [11]:

$$\mathrm{AF}_E\ [1/\mathrm{m}] = \frac{E}{V_r} \tag{9.18}$$

$$\mathrm{AF}_H\ [\mathrm{S/m}] = \frac{H}{V_r} \tag{9.19}$$

where,

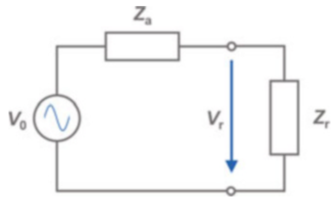


Fig. 9.6 Simplified equivalent circuit of a receiving antenna with open load circuit voltage $\underline{V}_0$ [V], antenna impedance $\underline{Z}_a$ [Ω], receiver impedance $\underline{Z}_r$ [Ω], and voltage at the receiving antenna terminals $\underline{V}_r$ [V]

E = RMS value of the electric field strength at the receiving antenna in [V/m]
H = RMS value of the magnetic field strength at the receiving antenna in [S/m]
V_r = RMS value of the voltage at the receiving antenna's terminals in [V]

In case of matched impedances at the receiving antenna's terminals (Fig. 9.6), free space, a plane wave, and aligned receiving antenna to the polarization of the incident wave, we can write:

$$P_r = S_r \cdot A_{em} \tag{9.20}$$

$$\frac{V_r^2}{\mathrm{Re}(\underline{Z}_r)} = \frac{E^2}{\eta} \cdot \frac{\lambda^2}{4\pi} G_{ir} \tag{9.21}$$

In case of free space and a plane wave, we can set $\lambda = c/f$ and $\eta = \eta_0 = 120\pi\ \Omega$ in Eq. 9.21 and get:

$$\mathrm{AF}_E^2 = \frac{E^2}{V_r^2} = \frac{\eta}{\mathrm{Re}(\underline{Z}_r)} \cdot \frac{4\pi}{\lambda^2} \cdot \frac{1}{G_{ir}} \tag{9.22}$$

$$\mathrm{AF}_E = \frac{E}{V_r} = \sqrt{\frac{480\pi^2}{c^2}} \cdot \frac{f}{\sqrt{\mathrm{Re}(\underline{Z}_r) G_{ir}}} \tag{9.23}$$

where,

P_r = RMS output power at the receiving antennas terminals in [W]
S_r = power density S_r at the receiving antenna in [W/m^2]
A_{em} = maximum effective aperture of the receiving antenna (which requires alignment of the wave and the receiving antenna polarization) in [m^2]
V_r = RMS voltage at the receiving antenna terminals in [V]
$\mathrm{Re}(\underline{Z}_r)$ = real value of the impedance of the receiver at the receiving antenna's terminals in [Ω]
η = intrinsic impedance (in the far-field: Z_w is equally to the intrinsic impedance η) of the incident plane wave at the receiving antenna in [Ω]
E = RMS value of the electric field strength at the receiving antenna in [V/m]

λ = wavelength of the sinusoidal incident plane wave in [m]
G_{ir} = antenna gain of the receiving antenna (dimensionless)

In case of $Z_r = 50\,\Omega$, we can write:

$$\mathrm{AF}_E\ [1/\mathrm{m}] = \frac{9.73}{\lambda\sqrt{G_{ir}}} \tag{9.24}$$

where,

AF_E = antenna factor of an E-field antenna in case of a plane wave (far-field) in free-space, a $50\,\Omega$-system and matched polarization of the incident E-field and the antenna in [1/m]
λ = wavelength of the sinusoidal incident plane wave in [m]
G_{ir} = antenna gain of the receiving antenna (dimensionless)

The antenna factor AF is often given in decibel:

$$\mathrm{AF}_E\ [\mathrm{dB}(1/\mathrm{m})] = 20\log_{10}(\mathrm{AF}_E\ [1/\mathrm{m}]) \tag{9.25}$$

$$\mathrm{AF}_H\ [\mathrm{dB}(\mathrm{S/m})] = 20\log_{10}(\mathrm{AF}_H\ [\mathrm{S/m}]) \tag{9.26}$$

9.7 Ground Plane Reflection and Multipath Effect

Assuming a transmitting and a receiving antenna placed above a reflective ground plane, where the receiving antenna and the ground plane are in the *far-field* of the transmitting antenna (see Fig. 9.7). The received field at the receiver antenna is the sum of the direct wave $\vec{E}_d$ [V/m] and the reflected wave $\vec{E}_r$ [V/m]:

$$\vec{E}_{sum} = \vec{E}_d + \vec{E}_r \tag{9.27}$$

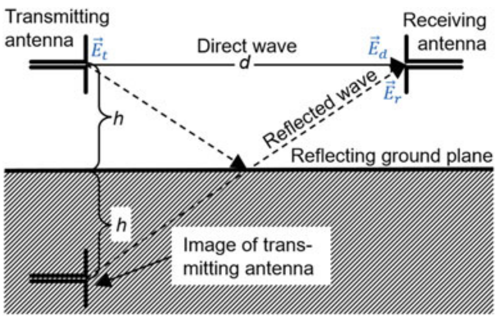


Fig. 9.7 Ground plane and receiving antenna are in the far-field of the transmitter

where,

$\vec{E}_{sum}$ = E-field sum of direct and reflected waves at receiving antenna in [V/m]
$\vec{E}_d$ = E-field at the receiving antenna from direct path in [V/m]
$\vec{E}_r$ = E-field at the receiving antenna from reflected path in [V/m]

This phenomena is called *multipath effect*. The field strength $|E_{sum}|$ [V/m] at the receiving antenna depends on the distances the waves have traveled, the polarization of the antennas, the reflection coefficient $\underline{\Gamma}$ of the ground plane, the attenuation α [1/m] of the media, and the frequency f [Hz] of the signal. Assuming that the ground plane and the receiving antenna are in the far-field of the transmitting antenna (so that R_{rad} is not influenced by the ground plane), a perfectly conducting ground plane ($\underline{\Gamma} = +1$ or -1 [2]) and a lossless media ($\alpha = 0$), the reflective field strength $|E_r|$ [V/m] could be as high as the field strength of the direct wave $|E_d|$ [V/m]. However, it could also happen that both fields cancel each other out at the receiving antenna. This means:

$$|E_{sum,max}| \approx 2|E_d| \tag{9.28}$$

$$|E_{sum,min}| \approx 0\,\text{V/m} \tag{9.29}$$

9.8 Intended Antennas

In EMC, the term *intended antenna* denotes an antenna for generating or receiving a specific electric, magnetic, or electromagnetic field in a predefined setup.

9.8.1 E-Field Antennas and Emission Testing

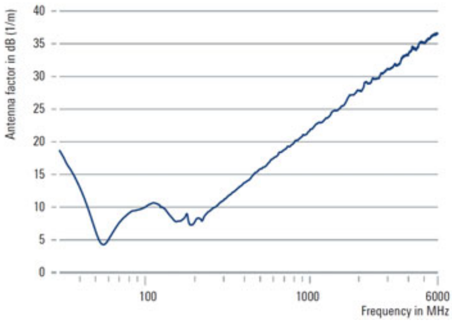
Radiated emission EMC standards like CISPR 11 [8], CISPR 32 [3], or FCC Part 15 [4] specify the radiated emission limits in [μV/m] or [dBμV/m]. Therefore, *E-field antennas* are used as receiving antennas (Fig. 9.8). In other words, the physical quantity of interest is the E-field [dBμV/m], and the measured physical quantity is the voltage $V_{measure}$ [dBμV] at the measurement equipment (e.g., spectrum analyzer or EMI receiver). The field strength of the E-field at the antenna can be calculated like this:

$$E = \text{AF}_E + \text{losses} + V_{measure} \tag{9.30}$$

where,

E = RMS field strength of the E-field at the antenna in [dBμV/m]
AF_E = antenna factor of an E-field antenna in [dB(1/m)]

(a)



(b)

Fig. 9.8 Example of an EMC emission and immunity testing E-field antenna. (**a**) Wideband biconilog antenna R&S®HL562E and frequency range from 30 MHz to 6 GHz. (**b**) Antenna factor of the wideband biconilog antenna R&S®HL562E by Rohde & Schwarz

Losses = losses of all cables and connectors from the receiver antenna to the measurement equipment (e.g., spectrum analyzer) in [dB]

$V_{measure}$ = measured RMS voltage at the measurement equipment (e.g., spectrum analyzer) in [dBμV]

9.8.2 E-Field Antennas and Immunity Testing

The physical quantity of interest for many *radiated immunity* EMC standards (like IEC 61000-4-3 [5]) is the E-field at a certain distance d (e.g., $E = 10\,\text{V/m}$ at $d = 10\,\text{m}$). Therefore, *E-field antennas* like bilogical, log-periodic, or horn antennas are used—basically the identical antennas as for E-field emission testing.

From the antenna fundamentals from above, we know how to calculate the maximum power density S_{max} [W/m^2] at a distance d [m] for a transmitting antenna in free space with antenna gain G_{it} and power P_t [W] at the transmitting antenna terminals:

$$S_{max} = \frac{\text{EIRP}}{4\pi d^2} = \frac{P_t G_{it}}{4\pi d^2} \tag{9.31}$$

where,

S_{max} = maximum power density for the input power of P_t to the antenna at distance d (in case of matched polarization, main lobe, free space, and line of sight in the far-field) in [W/m^2]

EIRP = RMS input power required to lossless isotropic radiator to give the identical maximum power density far from the antenna as the actual antenna in [W]

d = distance from the transmitting antenna (in the far-field and in the direction of the antenna's main lobe) in [m]
P_t = RMS power input to the transmitting antenna in [W]
G_{it} = antenna gain of the transmitting antenna compared to an isotropic radiator (dimensionless)

Furthermore, we know the power density S [W/m^2] of a plane wave in free space (far-field: wave impedance Z_w [Ω] is equal characteristic impedance $\eta_0 = 120\pi\ \Omega$) is given as [11]:

$$S = \frac{E^2}{\eta_0} = \frac{E^2}{120\pi\ \Omega} \tag{9.32}$$

where,

E = RMS value of the electric field strength in [V/m]
$\eta_0 = 120\pi\ \Omega$ = intrinsic impedance of free-space in [Ω]

If we combine Eqs. 9.31 and 9.32, we can determine the field strength E [V/m] at a given distance d [m] from the antenna for a given transmitting antenna input power P_t [W]:

$$E = \frac{\sqrt{30 P_t G_{it}}}{d} = \frac{\sqrt{30 P_t 10^{\frac{G_{it}[\mathrm{dBi}]}{10}}}}{d} \tag{9.33}$$

where,

E = RMS field strength of the electric field at distance d from the transmitting antenna for input power P_t (assuming free space, line of sight, distance d in the far-field and in the direction of the transmitting antenna's main lobe and matched polarization to the transmitting antenna) in [V/m]
P_t = RMS power input to the transmitting antenna in [W]
G_{it} = antenna gain of the transmitting antenna compared to an isotropic radiator (dimensionless)
G_{it}[dBi] = antenna gain of the transmitting antenna compared to an isotropic radiator in [dBi]
d = distance from the transmitting antenna in [m]

Thus, the required input power P_t [W] of a transmitting antenna for achieving a desired field strength E [V/m] at a given distance d [m] is (in free space, no reflective ground plane):

$$P_t = \frac{(E \cdot d)^2}{30 \cdot G_{it}} = \frac{(E \cdot d)^2}{30 \cdot 10^{\frac{G_{it}[\mathrm{dBi}]}{10}}} \tag{9.34}$$

where,

P_t = RMS input power at the transmitting antenna's input terminals for achieving electric field strength E at distance d (assuming free space, line of sight, distance d in the far-field and in the direction of the transmitting antenna's main lobe and matched polarization to the transmitting antenna) in [W]
E = RMS field strength of the electric field in [V/m]
G_{it} = antenna gain of the transmitting antenna compared to an isotropic radiator (dimensionless)
G_{it}[dBi] = antenna gain of the transmitting antenna compared to an isotropic radiator in [dBi]
d = distance from the transmitting antenna in [m]

9.8.3 H-Field Antennas and Emission Testing

Magnetic *radiated emission* EMC standards like MIL-STD-461 (RE101, $f = 30\,\text{Hz}$ to $100\,\text{kHz}$) require *H-field loop antennas* as receiving antennas (Fig. 9.9). In case the antenna factor is given in [dB(S/m)], the field strength of the H-field at the antenna can be calculated like this:

$$H = \mathrm{AF}_H + \text{losses} + V_{measure} \tag{9.35}$$

where,

H = RMS field strength of the H-field at the antenna in [dBμA/m]
AF_H = antenna factor of an H-field antenna in [dB(S/m)]
Losses = losses of all cables and connectors from the receiver antenna to the measurement equipment (e.g., spectrum analyzer) in [dB]
$V_{measure}$ = measured RMS voltage at the measurement equipment (e.g., spectrum analyzer) in [dBμV]

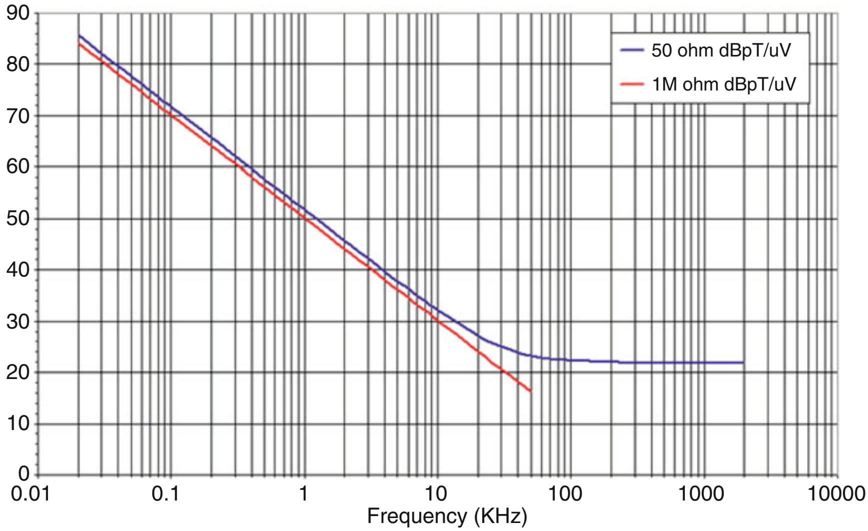
In case the antenna factor for the receiving loop antenna is given in [dBpT/μV], the field strength of the H-field at the antenna can be calculated like this:

$$H = (\mathrm{AF}_H + \text{losses}) \cdot V_{measure} \tag{9.36}$$

where,

H = RMS field strength of the H-field at the antenna in [dBpT]
AF_H = antenna factor of an H-field antenna in [dBpT/μV]
Losses = losses of all cables and connectors from the receiver antenna to the measurement equipment (e.g., spectrum analyzer) in [dB]
$V_{measure}$ = measured RMS voltage at the measurement equipment (e.g., spectrum analyzer) in [μV]

(a)



(b)

Fig. 9.9 Example of an emission testing H-field antenna. (**a**) A.H. Systems Inc. SA-560 passive loop antenna. (**b**) Coil factor of the passive loop antenna SAS-560 by A.H. Systems. The passive loop antenna SAS-560 was specifically designed for tests according to MIL-STD-461 RE101

9.8.4 H-Field Antennas and Immunity Testing

Radiated magnetic field EMC immunity standards, like IEC 61000-4-8 [7] (f = 50 Hz and 60 Hz) or MIL-STD-461 (RS101, f = 30 Hz to 100 kHz) require *induction coils* as transmitting antennas. In contrast to the E-field EMC tests—where the identical antennas are used for immunity and emission tests—the H-field immunity tests specify dedicated antennas, which are different from the H-field emission measurements.

Two important properties of an induction coil for H-field immunity tests are:

- **Homogeneity.** Magnetic field immunity tests often require that the field strength should not vary more than, e.g., ±3 dB (see Fig. 9.10) over the volume of the EUT. The magnetic field homogeneity over a given volume is dependent on the induction coil. Details about the induction coils and the test setups are given in the respective EMC standards.
- **Induction coil factor.** The induction *coil factor* is the antenna factor of an induction coil. It describes the ratio between the magnetic field strength H [A/m]

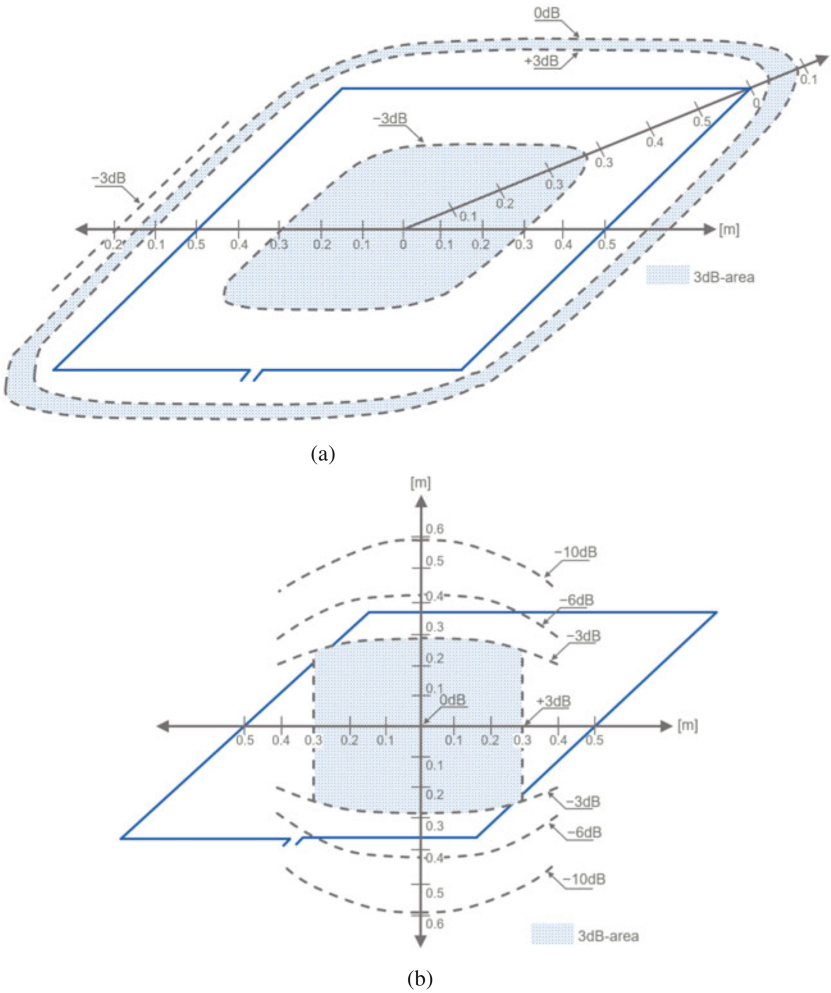


Fig. 9.10 H-field homogeneity of a 1×1 m square loop antenna [7]. (**a**) 3 dB—area in the plane to the loop antenna. (**b**) 3 dB—area orthogonal to the loop antenna

generated by an induction coil of given dimensions and the corresponding current I [A] value. The field is measured at the center of the coil plane without the EUT.

The manufacturer of an induction coil for magnetic immunity tests specifies in the datasheet of the coil antenna:

- The test volume and position of the homogeneous magnetic field region
- The operation frequency (range) of the antenna
- The coil factor

The magnetic field strength H [A/m] at the position of interest (relative to the induction coil) can be calculated like this:

$$H = c_f \cdot I \tag{9.37}$$

where,

H = RMS value of the magnetic field in the area of interest in [A/m]
c_f = coil factor of the induction coil for a given frequency (range) f and a predefined volume (at a defined relative position to the induction coil) in [(A/m)/A]
I = RMS value of the current flowing through the induction coil in [A]

9.9 Unintended Antennas

Current loops, PCB structures, cables, or slots may act as *unintended antennas*. As a result, they receive or emit electromagnetic radiation unintentionally.

9.9.1 E-Field from Differential-Mode Currents in Small Loops

Let us assume a setup like shown in Fig. 9.11 where the following is given [11]:

- Electrically small current loop (circumference $< \lambda/4$), so that the current distribution (magnitude, phase) is constant along the current loop.
- Measurement point of the E-field is in the far-field.

Given the points above, the maximum electric field $E_{DM,max}$ [V/m] caused by a differential-mode current $\underline{I}_{DM}$ [A] can be approximated as [11]:

$$E_{DM,max} \approx \begin{cases} 1.32 \cdot 10^{-14} \frac{I_{DM} f^2 A}{d} & \text{without reflecting ground plane} \\ 2.64 \cdot 10^{-14} \frac{I_{DM} f^2 A}{d} & \text{with reflecting ground plane (see Fig. 9.7)} \end{cases} \tag{9.38}$$

where,

$E_{DM,max}$ = maximum RMS field strength radiated by electrically small current loop with area A and differential current I_{DM} in [V/m]
I_{DM} = RMS value of the differential-mode current in [A]
f = frequency of the sinusoidal current signal in [Hz]
A = area of the current loop in [m^2]
d = distance to the center of the current loop in [m]

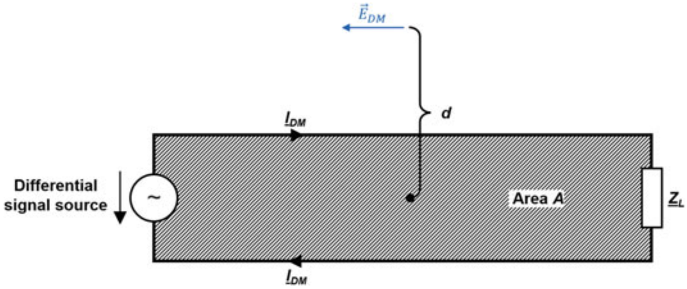


Fig. 9.11 Radiated emissions E_{DM} [V/m] due to differential-mode current I_{DM} [A] around a current loop with area A [m^2]. The direction of the electric field E_{DM} [V/m] is parallel to the wires

Reducing unintended radiation emissions due to a small differential-mode current loop gives the following options:

1. **Reducing the current amplitude.** It is usually not practical to reduce the amplitude of the signal current $\underline{I}_{DM}$ [A] because the amplitude was initially set due to functional reasons.
2. **Reducing the frequency or the amplitude of high-frequency harmonics.** Let us assume the signal is a digital signal. Because the radiated emissions are proportional to the square of the frequency f^2 [Hz], amplitudes of the higher-frequency harmonics of a digital signal should be kept low by increasing the rise /fall time of the digital signal. Section 5.3 describes how to reduce the amplitude of the high-frequency harmonics of a digital signal.
3. **Reducing the current loop area.** Loop areas can often not be changed without an extensive redesign. Therefore, loop areas should be considered early in the design phase.

Differential-mode currents do not radiate efficiently, and in practice, radiated emissions due to differential-mode currents are seldom an issue. Usually, the common-mode currents along cables are the root cause for high radiated emissions.

9.9.2 E-Field from Common-Mode Currents in Short Cables

Let us assume a setup like shown in Fig. 9.12 where the following is given [11]:

- Electrically short cable length l [m] (e.g., $l < \lambda/4$ [10]), so that the current distribution (magnitude, phase) is constant along the conductor.
- Measurement point of the E-field is in the far-field.

Given the points above, the maximum electric field $E_{CM,max}$ [V/m] caused by the common-mode current $\underline{I}_{CM}$ [A] can be approximated as [11]:

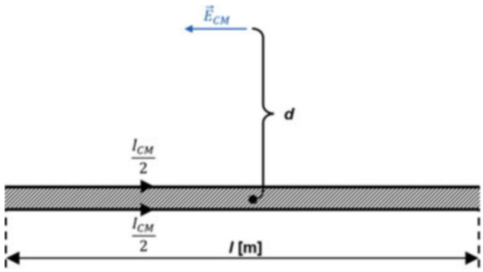


Fig. 9.12 Radiated emissions E_{CM} [V/m] due to common-mode current I_{CM} [A] through two closely spaced parallel conductors of length $l < \lambda/4$. The direction of the electric field E_{CM} [V/m] is parallel to the wires

$$E_{CM,max} \approx \begin{cases} 0.63 \cdot 10^{-6} \frac{I_{CM} f l}{d} & \text{without reflecting ground plane} \\ 1.26 \cdot 10^{-6} \frac{I_{CM} f l}{d} & \text{with reflecting ground plane (see Fig. 9.7)} \end{cases} \tag{9.39}$$

where,

$E_{CM,max}$ = maximum RMS field-strength radiated by electrically short cable of length l and common-mode current I_{CM} in [V/m]
I_{CM} = RMS value of the common-mode current in [A]
f = frequency of the sinusoidal current signal in [Hz]
l = length of the conductor or transmission line in [m]
d = distance to the center of the cable in [m]

Because common-mode currents are normally not required for system operation, the first rule is to avoid them. Once the common-mode currents are here: reduce them. One of the most common options to reduce common-mode currents is adding a clamp ferrite over the cable or wires, which are causing unintended radiated emissions. More information about cable mount ferrite beads can be found in Sect. 11.5.1 on page 169.

9.9.3 *Maximum E-Field from Common-Mode Signals*

Section 9.9.2 does only consider short cables (antennas) compared to the wavelength λ [m]. Therefore, these cables are not very effective radiators, which does not mean that they will never lead to failed EMC emission tests. Nonetheless, things get even worse when cable lengths l [m] match a multiple of a quarter wavelength $n\lambda/4$ [m]. This section presents simple approximations for the worst-case emissions from given common-mode voltages and currents.

9.9.3.1 Generic Calculation of Maximum E-Field

Let us assume a sinusoidal signal with a RMS voltage V_{RMS} [V] that drives a resonant structure like a $\lambda/4$-monopole antenna. As a result, the current distribution along the antenna has a RMS value of I_{RMS} [A]. Assuming a lossless antenna ($R_{loss} = 0\,\Omega$), where the antenna input impedance $\underline{Z}_{in}$ [Ω] is:

$$\underline{Z}_{in} = R_{in} + jX_{in} = R_{loss} + R_{rad} + jX_{in} = R_{rad} + jX_{in}, \tag{9.40}$$

we can calculate the radiated power P_{rad} [W] of this lossless antenna as [11]:

$$P_{rad} = \frac{V_{RMS}^2}{R_{rad}} = I_{RMS}^2 R_{rad} \tag{9.41}$$

Combining Eqs. 9.34 and 9.41 for a lossless transmitter (where directivity D is equal the antenna gain G_i), we get:

$$P_{rad} = \frac{V_{RMS}^2}{R_{rad}} = I_{RMS}^2 R_{rad} = \frac{E^2}{\eta_0} \frac{4\pi d^2}{G_i} \tag{9.42}$$

This leads us to the maximum radiated field-strength E_{max} [V/m] at the resonance of a lossless antenna with matched impedance, matched polarization, and line-of-sight path:

$$E_{max} \approx \sqrt{\frac{V_{RMS}^2 \cdot \eta_0 \cdot G_i}{R_{rad} \cdot 4\pi d^2}} = \sqrt{\frac{I_{RMS}^2 \cdot R_{rad} \cdot \eta_0 \cdot G_i}{4\pi d^2}} \tag{9.43}$$

where,

E_{max} = maximum E-field at distance d for matched impedances, matched polarization, line-of-sight path, and a lossless antenna driven by V_{RMS} in [V/m]
V_{RMS} = RMS voltage that drives the lossless antenna in [V]
I_{RMS} = RMS value of the current distribution along the lossless antenna in [A]
R_{rad} = radiation resistance of the lossless antenna (function of frequency) in [Ω]
$\eta_0 = 120\pi\ \Omega$ = intrinsic impedance of free space in [Ω]
$G_i = D$ = antenna gain of the lossless antenna (dimensionless)
d = distance from the antenna, where E_{max} is measured in [m]

9.9.3.2 Maximum E-Field for a Given Common-Mode Current

Assuming a common-mode current $\underline{I}_{CM}$ [A] flows along a resonant structure (e.g., a cable). When in resonance (and only then!), the maximum electric field-strength E_{max} [V/m] can be approximated based on Eq. 9.43 as:

$$E_{max} \approx \sqrt{\frac{R_{rad} \cdot \eta_0 \cdot G_i}{4\pi}} \frac{I_{CM}}{d} \tag{9.44}$$

$$E_{max} \approx \begin{cases} 60\frac{I_{CM}}{d} & \text{without reflecting ground plane} \\ 120\frac{I_{CM}}{d} & \text{with reflecting ground plane (see Fig.. 9.7)} \end{cases} \tag{9.45}$$

where,

E_{max} = maximum RMS E-field at distance d for matched impedances, matched polarization, line-of-sight path, and a resonant structure driven by I_{CM} in [V/m]
I_{CM} = RMS common-mode current that flows along the resonant structure in [A]
R_{rad} = radiation resistance in [Ω] (λ/2-dipole: 73 Ω, λ/4-monopole: 36.5 Ω)
$\eta_0 = 120\pi\ \Omega$ = intrinsic impedance of free-space in [Ω]
G_i = antenna gain (λ/2-dipole, 1.64; λ/4-monopole, 3.28)
d = distance from the antenna, where E_{max} is measured in [m]

9.9.3.3 Maximum E-Field for a Given Common-Mode Voltage

Assuming a common-mode voltage $\underline{V}_{CM}$ [V] that drives a resonant structure (e.g., a cable). When in resonance (and only then!), the maximum electric field-strength E_{max} [V/m] can be approximated based on Eq. 9.43 as:

$$E_{max} \approx \sqrt{\frac{\eta_0 \cdot G_i}{R_{rad} \cdot 4\pi}} \frac{V_{CM}}{d} \tag{9.46}$$

$$E_{max} \approx \begin{cases} 0.8\frac{V_{CM}}{d} & \lambda/2\text{-dipole without reflecting ground plane} \\ 1.6\frac{V_{CM}}{d} & \lambda/2\text{-dipole with reflecting ground plane (see Fig. 9.7),} \\ 1.6\frac{V_{CM}}{d} & \lambda/4\text{-monopole without reflecting ground plane} \\ 3.2\frac{V_{CM}}{d} & \lambda/4\text{-monopole with reflecting ground plane (see Fig. 9.7)} \end{cases} \tag{9.47}$$

where,

E_{max} = maximum RMS E-field at distance d for matched impedances, matched polarization, line-of-sight path, and a resonant structure driven by V_{CM} in [V/m]
V_{CM} = RMS common-mode voltage that drives the resonant structure in [V]
R_{rad} = radiation resistance in [Ω] (λ/2-dipole, 73 Ω; λ/4-monopole, 36.5 Ω)
$\eta_0 = 120\pi\ \Omega$ = intrinsic impedance of free space in [Ω]
G_i = antenna gain (λ/2-dipole, 1.64; λ/4-monopole, 3.28)
d = distance from the antenna, where E_{max} is measured in [m]

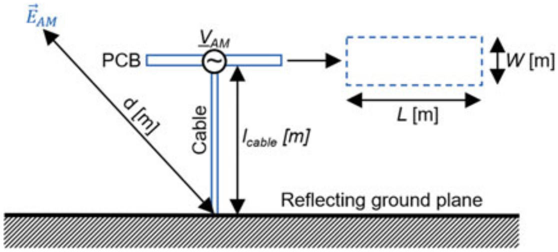


Fig. 9.13 Unintended antenna structure: cable connected to a PCBA on one end and attached to an infinite ground plane at the other end. An antenna-mode (common-mode) voltage $\underline{V}_{AM}$ [V] drives an antenna-mode current $\underline{I}_{AM}$ [A] through a cable

9.9.4 Maximum E-Field from PCB with Attached Cable

Let us assume a setup like shown in Fig. 9.13 where the following is given:

- A cable connected to a PCBA, which is driven by an unintended antenna-mode (common-mode) voltage $\underline{V}_{AM}$ [V] and connected to an infinite ground plane at the other end.
- The structure in Fig. 9.13 is essentially an unbalanced monopole antenna, which reaches the maximum radiation $E_{AM,max}$ [V/m] when the cable length l_{cable} [m] and/or the board size l_{board} [m] are a quarter wavelength $\lambda/4$ [m] or larger.
- Measurement point of the E-field is in the far-field, over a reflective ground plane.

The maximum RMS value of the electric field $E_{AM,max}$ [V/m] measured in a anechoic chamber (with a reflecting ground plane) caused by an antenna-mode signal with $\underline{V}_{AM}$ [V] and $\underline{I}_{AM}$ [A] that drives the board-cable structure as an antenna can be approximated like this [12–14]:

$$E_{AM,max} \approx \begin{cases} \frac{60 \cdot I_{AM,max}}{d} \cdot \frac{2}{\sin\left(\sqrt{2}\right)} & \text{when } l_{cable} \leq \frac{\lambda}{2} \\ \frac{60 \cdot I_{AM,max}}{d} \cdot \frac{2}{\sin\left(\sqrt{\frac{\lambda}{l_{cable}}}\right)} & \text{when } l_{cable} > \frac{\lambda}{2} \end{cases} \tag{9.48}$$

where,

$$I_{AM,max} = \frac{V_{AM}}{\frac{R_{min}}{k_{board} \cdot k_{cable}}} \tag{9.49}$$

and

$$k_{board} = \begin{cases} \sin\left(\frac{2\pi \cdot l_{board}}{\lambda}\right) & \text{when } l_{board} \leq \frac{\lambda}{4} \\ 1.0 & \text{otherwise} \end{cases} \tag{9.50}$$

$$k_{cable} = \begin{cases} \sin\left(\frac{2\pi \cdot l_{cable}}{\lambda}\right) & \text{when } l_{cable} \le \frac{\lambda}{4} \\ 1.0 & \text{otherwise} \end{cases} \tag{9.51}$$

$$l_{board} = \frac{1 + \frac{2L}{W}}{\frac{2L}{W}} \cdot \sqrt{L^2 + W^2} \tag{9.52}$$

V_{AM} = antenna-mode RMS voltage driving the monopole antenna (cable) in [V]
$I_{AM,max}$ = highest antenna-mode RMS current that exists on the cable in [A]
$R_{min} = 36.5\,\Omega$ = radiation resistance of a resonant $\lambda/4$-monopole antenna in [Ω]
l_{board} = effective board length in [m]
l_{cable} = cable length in [m]
L = length of the PCBA (board) in [m]
W = width of the PCBA (board) in [m]
k_{board} = impact of board size l_{board} on R_{min} in case $l_{board} \le \lambda/4$
k_{cable} = impact of cable length l_{cable} on R_{min} in case $l_{cable} \le \lambda/4$
d = distance to the radiating structure (PCBA, cable) in [m]
f = frequency of the sinusoidal signal $\underline{V}_{AM}$ in [Hz]
λ = wavelength of the sinusoidal signal $\underline{V}_{AM}$ in [m]
$c = 2.998 \cdot 10^8$ m/sec =speed of light

9.10 Free-Space Path Loss

The term free-space loss, free-space path loss factor, or *free-space path loss* (FSPL) refers to the attenuation of the electromagnetic field between a transmitting and a receiving antenna in case the space between the antennas is free of obstacles and a line-of-sight path through free-space.

A way to calculate the FSPL is given by the *Friis transmission equation*, which assumes matched impedances, matched polarization of the transmit, and receive antennas and that the receiving antenna is in the far-field [11]:

$$\frac{P_r}{P_t} = G_{it} G_{ir} \left(\frac{\lambda}{4\pi d}\right)^2 \tag{9.53}$$

$$\text{FSPL} = \left(\frac{4\pi d}{\lambda}\right)^2 \tag{9.54}$$

$$\text{FSPL [dB]} = 20\log_{10}\left(\frac{4\pi d}{\lambda}\right) = 20\log_{10}\left(\frac{4\pi d f}{c}\right) \tag{9.55}$$

where,

P_r = output power at the receiving antennas terminals in [W]
P_t = input power to the transmitting antenna in [W]
G_{ir} = antenna gain of the transmitting antenna (dimensionless)
G_{ir} = antenna gain of the receiving antenna (dimensionless)
λ = wavelength of the sinusoidal electromagnetic wave in [m]
d = distance between the transmitting and receiving antennas in [m]
f = frequency of the sinusoidal electromagnetic wave in [Hz]
$c = 1/(\sqrt{\mu_0 \epsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light

9.11 Link Budget

The *link budget* is an accounting of all of the power gains and losses that a signal experiences along a transmission path (from transmitter to receiver). If you want to calculate the received power P_r [dBm] for a given transmitter-receiver setup (transmitter, frequency, antennas, distance, etc.), you calculate the link budget. In simple terms, this means:

Received power P_r [dBm] = Transmitted power P_t [dBm] + Gains [dB] - Losses [dB]

If we rewrite the link budget formula above in a little more detailed way, we get:

$$P_r = P_t - L_t + G_{it} - \text{FSPL} - L_{misc} + G_{ir} - L_r \tag{9.56}$$

where,

P_r = maximum received power, e.g., at a spectrum analyzer (connected via coaxial cable with the receiving antenna) in case of the receiving antenna in the far-field, in the direction of the main lobe of the transmitting antenna, matched impedances, matched polarization of the incident wave, and the receiving antenna and free space in [dBm]
P_t = output power of, e.g., a signal generator (connected via coaxial cable with the transmitting antenna) in [dBm]
G_{ir} = antenna gain of the transmitting antenna in [dBi]
G_{ir} = antenna gain of the receiving antenna in [dBi]
L_t = transmitter losses (coaxial cable, connectors, etc.) in [dB]
FPSL = free-space path loss in [dB]
L_{misc} = miscellaneous losses (fading, polarization mismatch, etc.) in [dB]
L_r = receiver losses (coaxial cable, connectors, etc.) in [dB]

9.12 Summary

- **Antennas Types and EMC** There are dedicated antennas for the near-field and the far-fields and for the E- and H-fields (Table 9.1).

Table 9.1 Antenna types and EMC

	***E*-field**	***H*-field**
Near-field	E-field near-field probe	H-field near-field probe
Far-field	Dipole, log-periodic, horn, biconical, bilogical antenna	Loop antennas

- **Antenna directivity D.** The directivity D of an antenna is defined as the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions. The directivity does not take losses of the antenna into account.
- **Antenna gain G.** The antenna gain G corresponds to the directivity D of the antenna, but with losses (efficiency) of the antenna taken into account.
- **Antenna factor (AF).** Antennas for EMC tests are typically characterized by their antenna factors AF.
- **Unintended antennas.** Common-mode currents are more likely to lead to undesired radiation than differential-mode currents (compare Eqs. 9.38 vs. 9.39). From an EMC point of view, common-mode currents through unintended antennas should be prevented by design (radiated emissions are proportional to the common-mode current).
- **Free-space path loss (FSPL).** The FSPL is a practical and simplified approximation for the attenuation of an electromagnetic wave in the far-field (in the case of line of sight and free space).

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Chapter 10
Skin Effect

Nothing in life is to be feared, it is only to be understood. Now is the time to understand more, so that we may fear less.

—Marie Curie

10.1 Skin Depth

Introduction and definitions of the term skin depth δ [m] (Fig. 10.1):

- **Conductors.** The skin depth δ [m] is defined as the distance from the conductor surface where the current density has fallen (caused by the skin effect) to 37% = 1/e = 1/2.72 of the current density at the surface of the conductor J_0 [A/m^2]. The current density J_d [A/m^2] at distance d [m] from the conductor surface is defined as [2]:

$$|J_d| = |J_0| \cdot e^{-\frac{d}{\delta}} \tag{10.1}$$

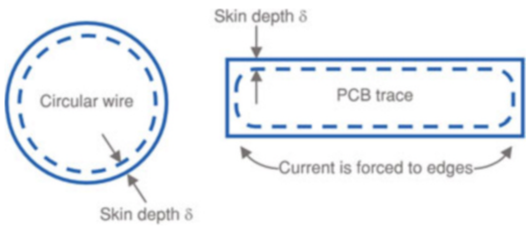
  where:

  $|J_d|$ = magnitude current density at distance d from the surface of the conductor in [A/m^2]
  $|J_0|$ = magnitude of the current density at the surface of the conductor in [A/m^2]
  d = distance from the conductor's surface in [m]
  δ = skin depth in [m]

- **Shielding.** Imagine an electromagnetic plane wave of field strength E_0 [V/m] and H_0 [A/m] entering an absorbing material (shield). The skin depth δ [m] is the distance an electromagnetic wave has to travel through that absorbing material until its field strength is reduced to 37% of E_0 or H_0 (Fig. 10.2). This means that the power of the plane electromagnetic wave is lowered by $10\log_{10}\left(1/e^2\right) = 8.686\,\mathrm{dB}$ after it has traveled the distance δ. The attenuation of an electromagnetic plane wave is defined like this [2]:

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_10

Fig. 10.1 Skin depth δ



$$|E_d| = \mathrm{Re}(|E_0| \cdot e^{-\underline{\gamma} d}) = \mathrm{Re}(|E_0| \cdot e^{-\alpha d} \cdot e^{-j\beta d}) = |E_0| \cdot e^{-\alpha d} = |E_0| \cdot e^{-\frac{d}{\delta}} \tag{10.2}$$

$$|H_d| = \mathrm{Re}(|H_0| \cdot e^{-\underline{\gamma} d}) = \mathrm{Re}(|H_0| \cdot e^{-\alpha d} \cdot e^{-j\beta d}) = |H_0| \cdot e^{-\alpha d} = |H_0| \cdot e^{-\frac{d}{\delta}} \tag{10.3}$$

where:

$|E_0|$ = electric field strength of an electromagnetic plane wave at the surface of a shield barrier, when entering that shield barrier in [V/m]
$|H_0|$ = magnetic field strength of an electromagnetic plane wave at the surface of a shield barrier, when entering that shield barrier in [A/m]
$|E_d|$ = electric field strength after the electromagnetic plane wave has traveled distance d through the shield barrier in [V/m]
$|H_d|$ = magnetic field strength after the electromagnetic plane wave has traveled distance d through the shield barrier in [A/m]
$\underline{\gamma}$ = the complex propagation constant of the shield barrier material [1/m]
α = attenuation constant of the shield barrier material in [1/m]
β = phase constant (or phase factor) of the shield barrier material in [rad/m]
d = distance from the shield barrier's surface in [m]
δ = skin depth in [m]

From Eqs. 10.2 and 10.3, we know that the skin depth δ [m] is defined as the inverse of the attenuation constant α [1/m] [1]:

$$\delta = \frac{1}{\alpha} = \frac{1}{\omega\sqrt{\frac{(\epsilon'\mu' - \epsilon''\mu'')}{2} \cdot \left(\sqrt{1 + \left(\frac{\epsilon'\mu'' + \epsilon''\mu'}{\epsilon'\mu' - \epsilon''\mu''}\right)^2} - 1\right)}} \tag{10.4}$$

where:

$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
ϵ' = real part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
ϵ'' = imaginary part of the complex permittivity ($\underline{\epsilon} = \epsilon' - j\epsilon''$) in [F/m]
μ' = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]
μ'' = imaginary part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]

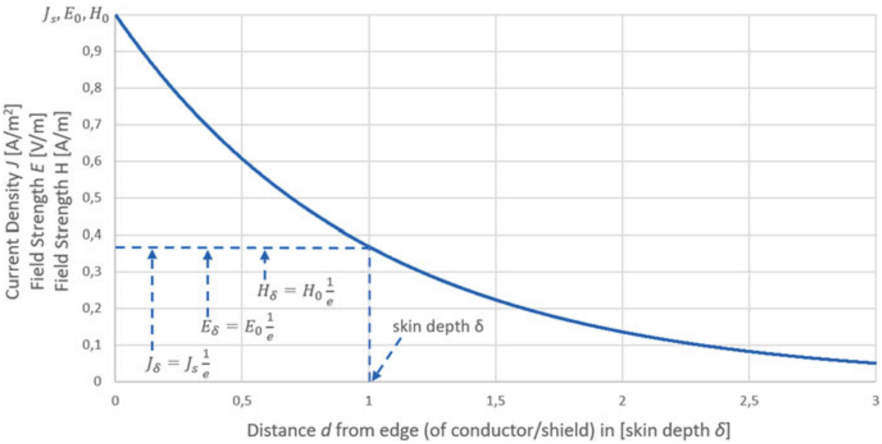


Fig. 10.2 Attenuation of current density J [A/m^2], electric field strength E [V/m], and magnetic field strength H [A/m] due to skin effect

For good conductors (with $\sigma \gg \omega\epsilon'$ and $\epsilon_r' = 1.0$) with negligible magnetic losses ($\mu'' = 0$), we can write [2]:

$$\delta = \sqrt{\frac{2}{\omega\mu_r'\mu_0\sigma}} = \sqrt{\frac{2}{\omega\mu_r'\mu_0\sigma}} = \frac{1}{\sqrt{\pi f \mu_r'\mu_0\sigma}} \tag{10.5}$$

where:

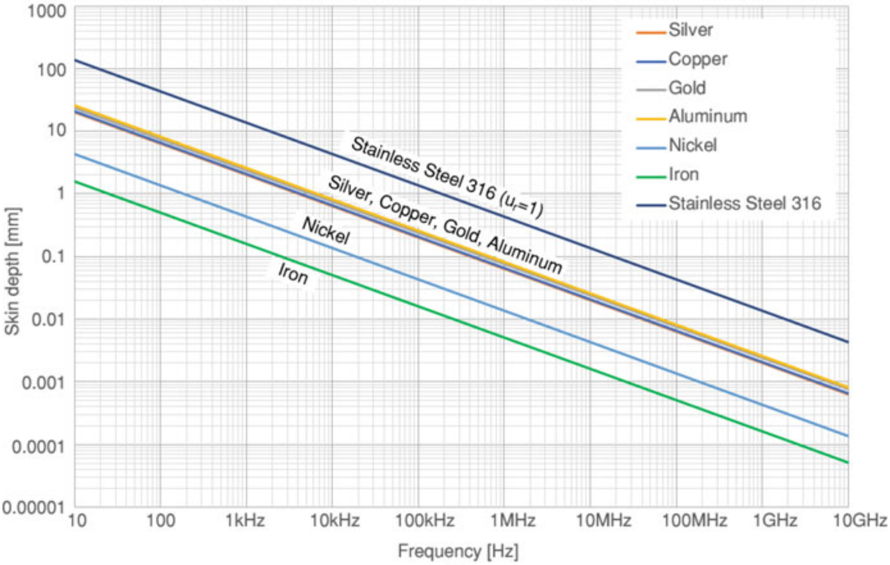
δ = skin depth in [m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
μ_r' = relative permeability of the material through which the signal current is flowing or through which the electromagnetic wave is traveling in [1]
$\mu_0 = 4\pi \cdot 10^{-1}$ H/m = 12.57 $\cdot 10^{-7}$ H/m = permeability of vacuum
σ = specific conductance of the material through which the signal current is flowing or through which the electromagnetic wave is traveling in [S/m]

Figure 10.3 presents some example values of skin depths for silver, copper, gold, aluminum, nickel, iron, and stainless steel 316. The following points have to be considered when calculating the skin depth for ferromagnetic metals (like nickel, iron):

- The relative permeability $\underline{\mu}_r$ depends on the specific material and alloy.
- Relative permeability $\underline{\mu}_r$ depends on the frequency f [Hz] and temperature T [K].

Parameters	Silver	Copper	Gold	Aluminum	Nickel	Iron	Stainless Steel 316
σ [S/m]	6,20E+07	5,82E+07	4,50E+07	3,80E+07	1,40E+07	1,00E+07	1,36E+06
μ_r [1]	1	1	1	1	100	1000	1
Frequency	Skin depth [mm]						
10 Hz	20,213	20,862	23,725	25,818	4,254	1,592	136,474
100 Hz	6,392	6,597	7,503	8,164	1,345	0,503	43,157
1 kHz	2,021	2,086	2,373	2,582	0,425	0,159	13,647
10 kHz	0,639	0,660	0,750	0,816	0,135	0,050	4,316
100 kHz	0,202	0,209	0,237	0,258	0,043	0,016	1,365
1 MHz	0,064	0,066	0,075	0,082	0,013	0,005	0,432
10 MHz	0,020	0,021	0,024	0,026	0,004	0,002	0,136
100 MHz	0,006	0,007	0,008	0,008	0,001	0,001	0,043
1 GHz	0,0020	0,0021	0,0024	0,0026	0,0004	0,0002	0,0136
10 GHz	0,0006	0,0007	0,0008	0,0008	0,0001	0,0001	0,0043

(a)



(b)

Fig. 10.3 Skin depth of some example metals. Permeability μ_r' is assumed to be constant. (**a**) Table with the assumed electrical conductivity σ and permeability μ_r'. (**b**) Skin depth δ as a function of frequency f

10.2 DC vs. AC Resistance

The resistance per-unit-length [Ω/m] for direct current (R'_{DC}) and for alternating current (R'_{AC}) of any conductor can be written as [2]:

$$R'_{DC} = \frac{\rho}{A} = \frac{1}{\sigma A} \tag{10.6}$$

$$R'_{AC} = \frac{\rho}{A_{eff}} = \frac{1}{\sigma A_{eff}} \tag{10.7}$$

where:

ρ = specific electrical resistivity of the conductor material in [Ωm]
σ = specific conductance of the conductor material in [S/m]
A = cross-sectional area of the conductor in [m^2]
A_{eff} = effective cross-sectional area of the conductor through which the current effectively flows in [m^2]

For direct current (DC, 0 Hz), the cross-sectional area A_{eff} [m^2] through which the DC current flows is equal to the conductor cross-sectional area A [m^2]. However, for high-frequency AC current with frequency f [Hz], the magnetic field—produced by current in the conductor—forces the current flow toward the outer surface of the conductor, and as a consequence of that, the current density increases exponentially from the core of the conductor toward the conductor's surface. The higher the signal frequency f [Hz], the smaller the cross section A_{eff} [m^2] through which the current effectively flows.

The accurate calculation of A_{eff} [m^2] is difficult but can be reasonably approximated by assuming that the current is uniformly distributed over the skin depth δ [m]. In case of $D \gg \delta$, the approximate $A_{effWire}$ [m^2] of a round conductor (circular wire) with diameter D [m] can be calculated as follows:

$$A_{effWire} \approx A - A_{noCurr} = \frac{D^2}{4}\pi - \frac{(D-2\delta)^2}{4}\pi = \left(D\delta - \delta^2\right)\pi \tag{10.8}$$

where:

A = cross-sectional area of the round conductor with diameter D in [m^2]
A_{noCurr} = the approximate cross-sectional area of the conductor where no current flows (as a consequence of the skin effect) in [m^2]
δ = skin depth in [m]
D = diameter of round conductor in [m]

For a PCB trace according to Fig. 10.4, the effective area $A_{effPCBtrace}$ [m^2] can be approximately calculated like this:

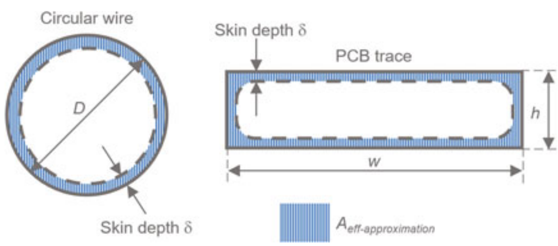


Fig. 10.4 Approximation of A_{eff} using the skin depth δ

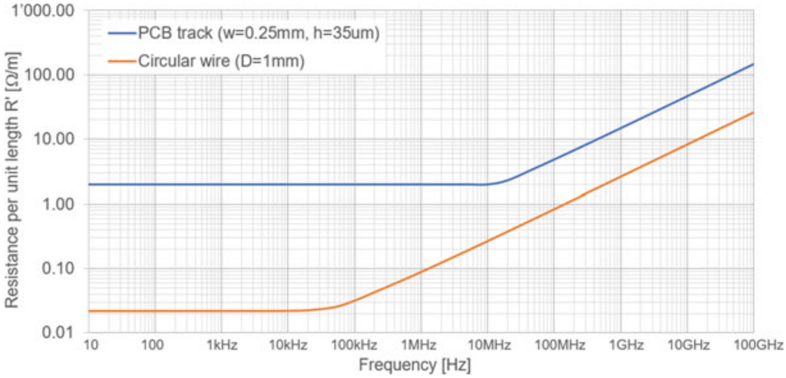


Fig. 10.5 Increased resistance due to skin effect: PCB copper trace (width $w = 0.25$ mm, height $h = 35\,\mu\text{m} = 1$ oz) vs. round copper wire (diameter $D = 1$ mm)

$$A_{effPCBtrace} \approx A - A_{noCurr} = wh - (w - 2\delta)(h - 2\delta) = 2\delta(w + h) - 4\delta^2 \tag{10.9}$$

where:

A = cross-sectional area of the PCB trace with width w and height h in [m^2]
A_{noCurr} = the approximate cross-sectional area of the conductor where no current flows (as a consequence of the skin effect) in [m^2]
δ = skin depth in [m]
w = width of the PCB trace in [m]
h = height or thickness of the PCB trace in [m]

Figure 10.5 shows R'_{AC} [Ω/m] of a PCB trace versus a round copper wire. The calculations for R'_{AC} [Ω/m] are approximations and ignore the return current path (proximity effect) and assume a single conductor surrounded by air only. However, the diagram gives an idea of how the skin effect influences the resistance at higher frequencies.

10.3 Surface Resistance

The *surface resistance* (or *sheet resistance*) R_s [Ω/square] for a thin layer with height h [m]—like shown in Fig. 10.6—is defined as [4]:

$$R_s = \frac{1}{\sigma h} \tag{10.10}$$

where:

σ = specific conductance of the conductor material in [S/m]

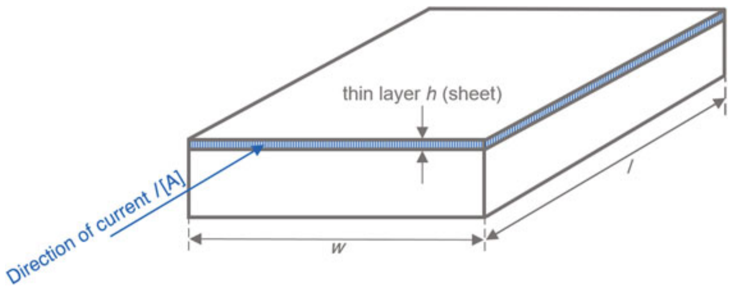


Fig. 10.6 Sheet (thin layer) of height of h, width w, and length l. Current flow is parallel to length l (not perpendicular to the sheet)

h = height or thickness of a thin conductor layer [m]

For a good conductor ($\sigma \gg \omega\epsilon'$), with negligible magnetic losses $\mu'' = 0$ and at high frequency (skin depth δ much smaller than conductor cross-sectional dimensions), the thickness of the conductor sheet h [m] can be set to the skin depth δ [m], and the surface resistance R_s [Ω/square] is defined as [3]:

$$R_s = \frac{1}{\sigma\delta} = \sqrt{\frac{\omega\mu'}{2\sigma}} \tag{10.11}$$

where:

σ = specific conductance of the conductor material in [S/m]
δ = skin depth in [m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\mu' = \mu_r' \mu_0$ = ability to store energy in a medium when an external magnetic field is applied = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]

At high frequencies, where the current flows at the surface of the conductor (skin effect), the resistance of a conductor with perimeter p [m] and length l [m] can be approximated as:

$$R \approx R_s \frac{l}{p} = \frac{l}{\sigma\delta p} \tag{10.12}$$

where:

R_s = surface resistance of the conductor material in [Ω/square]
l = length of the conductor in [m]
p = perimeter of the conductor in [m]
δ = skin depth in [m]

10.4 Summary

- **Skin Depth δ of Good Conductor.**

$$\delta = \frac{1}{\sqrt{\pi f \mu_r' \mu_0 \sigma}} \tag{10.13}$$

where:

μ_r' = relative permeability of the conductor material in [1]
$\mu_0 = 4\pi \cdot 10^{-1}$ H/m = 12.57 $\cdot 10^{-7}$ H/m = permeability of vacuum
σ = specific conductance of the conductor material in [S/m]

- **Resistance vs. Frequency.** Due to the skin effect, the resistance of a conductor increases with increasing frequency. As an approximation and in case the skin depth δ [m] is much smaller than the conductor's outer dimensions, it can be assumed that the current is uniformly distributed over the skin depth δ.
- **Surface Resistance**. Surface resistance R_s [Ω/square] of a good conductor at high frequency (skin depth δ much smaller than conductor cross-sectional dimensions):

$$R_s = \frac{1}{\sigma\delta} = \sqrt{\frac{\omega\mu'}{2\sigma}} \tag{10.14}$$

where:

σ = specific conductance of the conductor material in [S/m]
δ = skin depth in [m]
$\omega = 2\pi f$ = angular frequency of the signal in [rad/sec]
$\mu' = \mu_r'\mu_0$ = ability to store energy in a medium when an external magnetic field is applied = real part of the complex permeability ($\underline{\mu} = \mu' - j\mu''$) in [H/m]

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Chapter 11
Components

We will make electricity so cheap that only the rich will burn candles.

—Thomas A. Edison

11.1 Conductors

11.1.1 Definition of a Conductor

First, let us give a criterion for an electrical *conductor* and an insulator (dielectric material) [16]:

- **Conductor.** $\left(\sigma / \left(\omega\epsilon'\right)\right)^2 \gg 1.0$
- **Dielectric.** $\left(\sigma / \left(\omega\epsilon'\right)\right)^2 \ll 1.0$

where:

$\sigma =$ specific conductance of the material in [S/m]
$\omega = 2\pi f =$ angular frequency of the sinusoidal signal in [rad/sec]
$\epsilon' = \epsilon_r' \epsilon_0 =$ permittivity of the material in [F/m]

11.1.2 Conductor Equivalent Circuits

Figure 11.1 shows that the equivalent circuit of a single conductor is a simple RL-circuit, where the self-inductance L [H] is split into an internal and an external self-inductance:

$$L = L_{int} + L_{ext} = \frac{\Phi_{int}(t)}{i(t)} + \frac{\Phi_{ext}(t)}{i(t)} \tag{11.1}$$

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_11

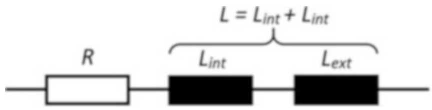


Fig. 11.1 Equivalent lumped-element circuit of a single conductor (no capacitance, because return current path and other parallel conductors are not considered)

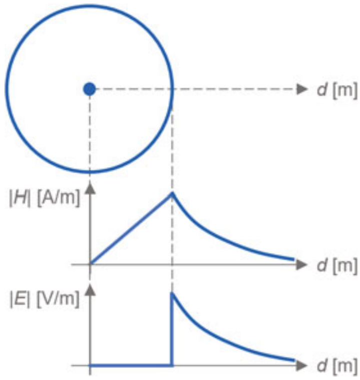


Fig. 11.2 Internal and external E- and H-field of a single round conductor with uniform current distribution (DC) in a homogeneous environment

where:

L = total partial inductance of a single conductor in [H]
L_{int} = internal inductance of a single conductor in [H]
L_{ext} = external inductance of a single conductor in [H]
$\Phi_{int}(t)$ = internal magnetic flux of conductor caused by the current I in [Wb]
$\Phi_{ext}(t)$ = external magnetic flux of conductor caused by the current I in [Wb]
$i(t)$ = current through the conductor in [A]

Figure 11.2 shows that a single conductor has an internal H-field [A/m], but the internal E-field [V/m] is zero (nonexistent). Thus, a single conductor has an internal inductance L_{int} [H] (linked to the internal H-field) but no internal capacitance (because of the missing E-field). In general, the internal inductance L_{int} [H] is much smaller than the external inductance L_{ext} [H]. This is especially valid for high frequencies (more details in Sect. 11.1.4).

At increasing signal frequency f [Hz], a conductor changes its impedance $\underline{Z}(f)$ [Ω]. Sections 11.1.3–11.1.7 present the calculations of the high-frequency equivalent circuit parameters of the models shown in Fig. 11.1 (single conductor) and Fig. 11.3 (conductor and its return current path). Here is an explanation of the equivalent lumped-element circuit models:

- **Resistance.** R' [Ω/m] is the per-unit-length resistance and $R = R' \cdot l$ [Ω] the resistance of a conductor with a certain length l [m].
- **Internal self-inductance.** L'_{int} [H/m] is the per-unit-length internal self-inductance of a conductor and $L_{int} = L'_{int} \cdot l$ [H] the internal self-inductance of a conductor with a certain length l [m].

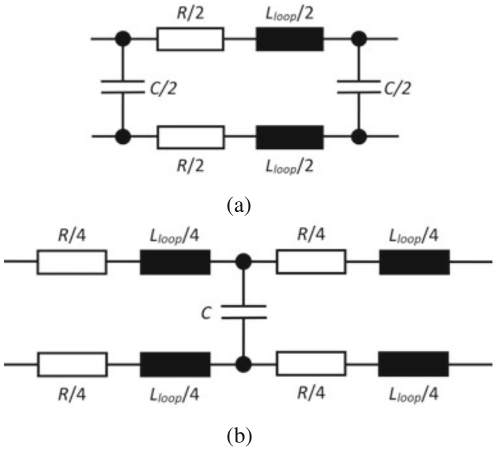


Fig. 11.3 Equivalent lumped-element circuit of a closed current loop (e.g., parallel conductors or PCB traces with forward and backward path). (**a**) Lumped-element π-model. (**b**) Lumped-element T-model

- **External self-inductance.** L'_{ext} [H/m] is the per-unit-length external self-inductance of a conductor and $L_{ext} = L'_{ext} \cdot l$ [H] the external self-inductance of a conductor with a certain length l [m].
- **Mutual inductance.** M'_{ij} [H/m] is the *mutual inductance* per-unit-length and $M_{ij} = M'_{ij} \cdot l = \Phi_{ij}/i_j(t)$ [H] the mutual inductance of the i-th current loop of length l [m]. Φ_{ij} is the magnetic flux through the i-th current loop, which is caused by the current $i_j(t)$ [A], which flows through the j-th current loop. Mutual inductance is present in the case of multiple conductors and current loops involved. The consequence of mutual inductance is that a change in current in the j-th current loop can induce a voltage in the nearby i-th current loop.
- **Loop inductance.** L'_{loop} [H/m] is the per-unit-length inductance of a current loop and $L_{loop} = L'_{loop} \cdot l$ [H] is the total inductance of a current loop: $L_{loop} = L_{total} = L_{int1} + L_{ext1} + L_{int2} + L_{ext2} - 2M_{12}$ (assuming $M_{12} = M_{21}$). Note: in case that the currents through two parallel conductors flow in the same direction (and not in the opposite direction like in the case of a current loop, which has a forward and return current flowing through two parallel conductors), the mutual inductances of these two conductors (M_{12} and M_{21}) have to be added and not subtracted and the total inductance is: $L_{total} = L_{int1} + L_{ext1} + L_{int2} + L_{ext2} + 2M_{12}$ [16].
- **Capacitance.** C' [F/m] is the per-unit-length capacitance and $C = C' \cdot l$ [F] the capacitance of a transmission line (consisting of a forward and return current path) with a certain length l [m].

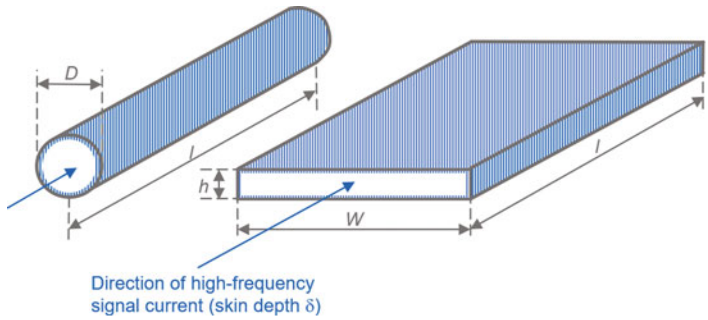


Fig. 11.4 Surface current at high frequency

11.1.3 Resistance of a Conductor

The *resistance* of a conductor depends primarily on the electrical conductivity σ [S/m] of the conductor's material and the signal frequency f [Hz] (skin effect; see Fig. 10.5).

In the case of DC or low frequencies, where the radius r [m] of a round conductor or the height h [m] of a PCB trace is much smaller than the skin depth δ [m], the resistance of a conductor can be calculated to:

$$R_{LF} = R'_{LF} \cdot l = \frac{1}{\sigma A} \cdot l, \text{ at low-frequency} \tag{11.2}$$

where:

R_{LF} = low-frequency (LF) resistance of a conductor of length l in [Ω]
σ = specific conductance of the conductor material in [S/m]
A = cross-sectional area of the conductor in [m^2]
l = length of the conductor in [m]

For high frequencies, the skin effect kicks in, and the current starts to flow close to the conductor's surface (see Figs. 11.4 and 11.5). Therefore, when the cross-sectional area—through which the current flows—is mainly determined by the skin depth δ [m], the resistance of the conductor can be approximately calculated as [2]:

$$R_{HF} \approx \frac{R_s}{p} \cdot l = \frac{l}{\sigma \delta p} = \frac{l}{p}\sqrt{\frac{\pi f \mu'}{\sigma}}, \text{ at high-frequency} \tag{11.3}$$

where:

R_{HF} = high-frequency (HF) resistance of a conductor of length l in [Ω]
$R_s = 1/(\sigma\delta)$ = surface resistance of the conductor material (read more about the term surface resistance in Sect. 10.3) in [Ω/square]
$\mu' = \mu_r \mu_0$ = permeability of the conductor in [H/m]

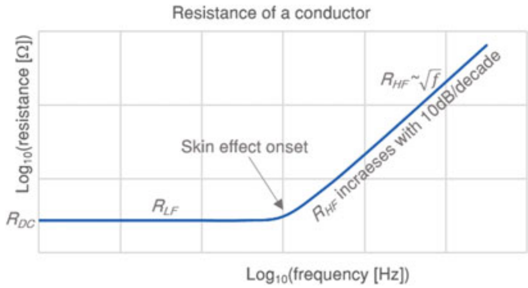


Fig. 11.5 Resistance of a conductor vs. frequency

f = frequency of the sinusoidal signal in [Hz]
p = perimeter/circumference of the cross section of the conductor in [m]
l = length of the conductor in [m]

For a circular wire ($p = D\pi$) or a rectangular conductor ($p = 2wh$), the high-frequency resistance can be approximated as:

$$R_{HF,round} \approx R_s \frac{l}{D\pi} = \frac{l}{\sigma\delta D\pi}, \text{ at high-frequency} \tag{11.4}$$

$$R_{HF,rectangular} \approx R_s \frac{l}{2wh} = \frac{l}{\sigma\delta 2wh}, \text{ at high-frequency} \tag{11.5}$$

where,

$R_{HF,round}$ = high-frequency (HF) resistance of a round conductor with diameter D [m] and length l [m] in [Ω]
$R_{HF,rectangular}$ = high-frequency (HF) resistance of a rectangular conductor with a of width w [m] and height h [m] in [Ω]
$R_s = 1/(\sigma\delta)$ = surface resistance of the conductor material in [Ω/square]
D = diameter of the round conductor in [m]
w = width of rectangular conductor in [m]
h = height or thickness of rectangular conductor in [m]

11.1.4 Internal Inductance of a Conductor

The *internal self-inductance* L_{int} [H] is present due to the magnetic flux Φ_{int} [Wb] through the conductor itself. The internal self-inductance is usually much smaller than the external inductance ($L_{int} \ll L_{ext}$), because normally the conductor diameter is smaller than the loop area of the magnetic flux outside the conductor. In addition, the internal inductance is frequency dependent and decreases by factor $\sqrt{f}$ (once the skin effect comes into play, see Fig. 11.6).

The internal self-inductance for DC and low-frequency $L_{int,LF}$ of round conductors (e.g., wires) and rectangular conductors (e.g., PCB traces with width w and

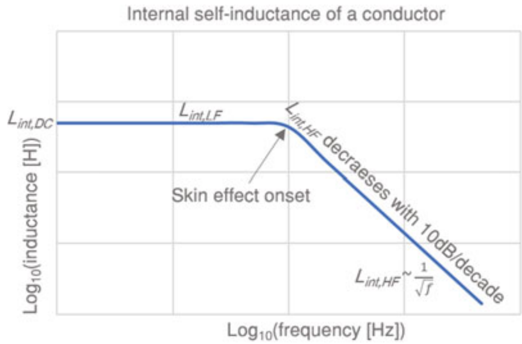


Fig. 11.6 Internal self-inductance of a conductor vs. frequency

thickness/height h) can be approximated as [9]:

$$L_{int,LF,round} \approx \frac{\mu'}{8\pi} \cdot l = 50\,\mathrm{nH/m} \cdot l, \text{ at low-frequency} \tag{11.6}$$

$$L_{int,LF,rectangular} \approx \frac{\mu'}{8} \frac{wh}{(w+h)^2} \cdot l, \text{ at low-frequency} \tag{11.7}$$

where:

$\mu' = \mu_r' \mu_0 = \mu_0 =$ permeability of the conductor material in [H/m]
$w =$ width of rectangular conductor in [m]
$h =$ height or thickness of rectangular conductor in [m]
$l =$ length of the conductor in [m]

Equation 11.6 shows that the internal inductance of a wire does not depend on the diameter D [m] of the wire. For rectangular conductors, there exist no closed-form solutions for the internal inductance [9].

The internal self-inductance for high frequency can be approximated as [2, 16]:

$$L_{int,HF} \approx \frac{R_s}{\omega p} \cdot l = \frac{l}{\omega \sigma \delta p} = \frac{R_{HF}}{\omega}, \text{ at high-frequency} \tag{11.8}$$

where:

$R_s = 1/(\sigma\delta)$ surface resistance of the conductor material in [Ω/square]
$R_{HF} =$ high-frequency (HF) resistance in [Ω]
$l =$ length of the conductor in [m]
$\omega = 2\pi f =$ angular frequency of the sinusoidal signal in [rad/sec]
$p =$ perimeter/circumference of the cross section of the conductor in [m]
$\sigma =$ specific conductance of the conductor material in [S/m]
$\delta =$ skin depth [m]

11.1.5 Internal Impedance of a Conductor

Every conductor has an *internal impedance* which can be modeled as a lumped-element RL-model (see Fig. 11.1) with resistance R and internal inductance L_{int}:

$$\underline{Z}_{int} = R + j\omega L_{int} = \left(R' + j\omega L'_{int}\right) \cdot l \tag{11.9}$$

where:

$R = R'l$ = electrical resistance of the conductor in [Ω]
$L_{int} = L'_{int}l$ = internal self-inductance of the conductor in [H]
$\omega = 2\pi f$ = angular frequency of the sinusoidal signal in [rad/sec]
l = length of the conductor in [m]
$j = \sqrt{-1}$ = imaginary unit

For DC and low frequencies (below skin effect), the internal impedance of a conductor is:

$$\underline{Z}_{int,LF} = R_{LF} + jL_{int,LF}, \text{ at low frequency} \tag{11.10}$$

where:

R_{LF} = LF resistance according to Eq. 11.2 in [Ω]
$L_{int,LF}$ = LF internal inductance according to Eqs. 11.6 and 11.7 in [H]
$j = \sqrt{-1}$ = imaginary unit

Figure 11.7 shows that once the skin effect comes into play, the following happens:

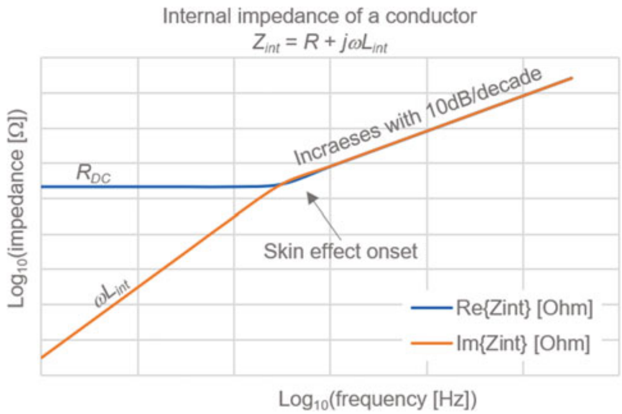


Fig. 11.7 Internal impedance of a conductor vs. frequency

- The resistance increases by the factor $\sqrt{f}$ and the internal self-inductance decreases by the factor $\sqrt{f}$.
- The imaginary part and the real part of the internal impedance are identical: $\text{Re}(\underline{Z}_{int,HF}) = R_{HF} = \text{Im}(\underline{Z}_{int,HF}) = \omega L_{int,HF}$.

Therefore, for the internal impedance of a conductor at high frequency, we can write:

$$\underline{Z}_{int,HF} = (1 + j)\, R_{HF} = (1 + j)\, \frac{l}{p}\sqrt{\frac{\pi f \mu'}{\sigma}}, \text{ at high-frequency} \tag{11.11}$$

where:

R_{HF} = HF resistance according to Eq. 11.3 in [Ω]
l = length of the conductor in [m]
p = perimeter/circumference of the cross section of the conductor in [m]
f = frequency of the sinusoidal signal in [Hz]
$\mu' = \mu_r \mu_0$ = permeability of the conductor in [H/m]
σ = specific conductance of the conductor material in [S/m]
$j = \sqrt{-1}$ = imaginary unit

11.1.6 External Inductance of a Single Conductor

The *external inductance* L_{ext} [H] is present due to the flux outside (external) of the conductor. It is to be mentioned that L_{ext} is much bigger than L_{int} because the conductor diameter is usually much smaller than the loop area of the magnetic flux.

The self-inductance of a single round conductor L_{round} [H] and a single rectangular conductor $L_{rectangular}$ [H], like shown in Fig. 11.8, in a surrounding with $\mu_r = 1$, can be approximated as [16]:

$$L_{round} \approx 2 \cdot 10^{-5} \cdot l \left[\ln\left(\frac{4l}{D}\right) - 1 + \frac{D}{2l} + \frac{\mu_r' T(x)}{4} \right] \tag{11.12}$$

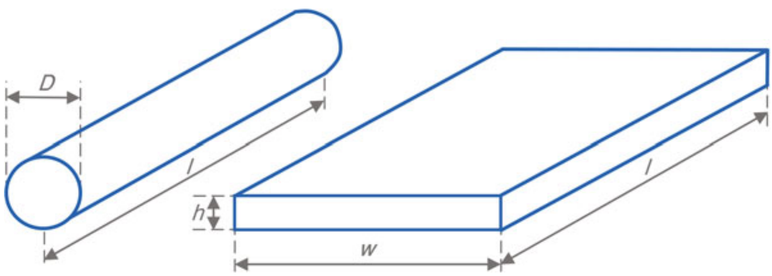


Fig. 11.8 Round conductor with diameter D (e.g., wire) and rectangular conductor with width w and height or thickness h

$$L_{rectangular} \approx 2 \cdot 10^{-5} \cdot l \left[\ln \left(\frac{2l}{w+h} \right) + 0.25049 + \frac{w+h}{3l} + \frac{\mu_r' T(x)}{4} \right] \tag{11.13}$$

where:

l = length of the round conductor in [m]
D = diameter of the round conductor in [m]
w = width of rectangular conductor in [m]
h = height or thickness of rectangular conductor in [m]
μ_r' = relative magnetic permeability of the round conductor
$T(x) = \sqrt{\frac{0.873011+0.00186128x}{1-0.278381x+0.127964x^2}}$ = compensation of the inductance for AC effects
$x = D\pi\sqrt{\frac{2\mu' f}{\sigma}}$
$\mu' = \mu_r' \mu_0$ = permeability of the conductor material in [H/m]
f = frequency of the sinusoidal signal in [Hz]
σ = specific conductance of the conductor material in [S/m]

Even more simplified (neglecting $T(x)$, $\mu_r = 1$), we can write [16]:

$$L_{round} \approx 2 \cdot 10^{-5} \cdot l \left[\ln \left(\frac{4l}{D} \right) - 0.75 \right] \tag{11.14}$$

$$L_{rectangular} \approx 2 \cdot 10^{-5} \cdot l \left[\ln \left(\frac{2l}{w+h} \right) + 0.5 + 0.2235 \left(\frac{w+h}{l} \right) \right] \tag{11.15}$$

Remember that these formulas above for L_{round} [H] and $L_{rectangular}$ [H] do only give you the *partial self-inductance* of a dedicated conductor with length l. When looking at a current loop, which consists of different conductor parts, all partial self-inductance and mutual inductances of all the conductor parts which eventually form the current loop have to be considered.

11.1.7 Conductor with Return Current Path

If you do not only consider a single conductor but also the return current path, we have a transmission line and the following happens:

- **Resistance.** At high frequencies—where currents flow primarily on the conductor surface,—the interaction between the forward and its return current leads to a nonuniform current distribution on the surface of conductors [9]. This effect is called the *proximity effect* and has a slight influence on the HF resistance R_{HF} [Ω]. The proximity effect will not be discussed further, and the proximity effect is ignored in all considerations throughout this book.

- **Inductance.** Currents flow in loops. Therefore, all the involved self-inductance and mutual-partial inductances of the forward and return current sum up to the *loop inductance* L_{loop} [H] as shown in Fig. 11.3.
- **Capacitance.** As soon as there is not just a single conductor, but also a return current path, there is also a capacitance C [F] to be considered. The equivalent RLC-circuit of such a setup is described in Fig. 11.3.

11.1.7.1 Conductor Length < ($\lambda/10$)

In case the length l [m] of the conductor is shorter than a tenth of the wavelength λ [m], the conductor and its return current path can be modeled with lumped RLC-elements (see Fig. 11.3). However, in case the signal line is longer than $\lambda/10$, reflections and ringing may occur and the signal line should not be modeled with a simple RLC-circuit. Instead, the characteristic impedance Z_0 [Ω] and the distributed parameter model are more appropriate to do so (see Sect. 7.3).

The calculations of the resistance R [Ω] and the per-unit-length resistance R' [Ω/m] of a conductor at low and high frequencies are presented in Sect. 11.1.3. The inductance L [H] and the capacitance C [F] of a conductor and its return current path depend on several parameters: the geometry of the conductor (and its return current path), the distance d [m] between the forward and backward current, the permeability $\mu' = \mu_r' \mu_0$ of the material(s) through which the magnetic flux flows, and the permittivity $\epsilon' = \epsilon_r' \epsilon_0$ of the material(s) through which the electric field lines are formed. Formulas for calculating the per-unit-length inductance L' [H/m] and capacitance C' [F/m] are presented in Sect. 7.4.

11.1.7.2 Conductor Length > ($\lambda/10$)

If the conductor length l [m] is longer than a tenth of the signal wavelength ($l > \lambda/10$), it is no longer enough to model the signal path as a simple lumped-element circuit with R [Ω], L [H], and C [F]. Instead, the conductor and its return current path should be modeled as transmission line according to the distributed parameter model (according to Fig. 7.2, with R' [Ω/m], L' [H/m], C' [F/m], and G' [S/m]) and with the characteristic impedance Z_0 [Ω].

Formulas for calculating the characteristic impedance Z_0 [Ω] of common transmission lines is presented in Sect. 7.3.

11.2 Resistors

There are, roughly said, three basic classes of passive *resistors*. All of them are designed for different applications and have their own advantages and drawbacks:

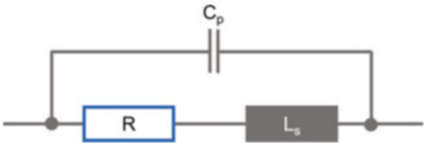


Fig. 11.9 Simplified high-frequency equivalent circuit of a resistor [3, 4]

- **Carbon Composition Resistors.** High energy surge applications. Negligible parasitic inductance L [14] and relatively high parasitic capacitance C [4].
- **Film-Type Resistors.** General-purpose low-power resistors (carbon, metal, metal oxide, thick and thin film). Low parasitic inductance L and capacitance C.
- **Wire-wound Resistors.** For high power applications. High parasitic inductance L and capacitance [4].

The impedance $\underline{Z}$ [Ω] of an ideal resistor is:

$$\underline{Z} = R \tag{11.16}$$

The simplified high-frequency equivalent circuit for resistors is presented in Fig. 11.9. The resistor's parasitic inductance L_s [H] and capacitance C_p [F] values depend on the resistor material type, the mechanical construction (SMT, axial leaded), and the nominal resistance value (low or high resistance). Be aware that the equivalent circuit model in Fig. 11.9 does not include the parasitic effects of the resistor's mounting parts (PCB pads). The impedance $\underline{Z}$ of a nonideal resistor can be written as:

$$\underline{Z} = \frac{R + j\omega L_s}{1 - \omega^2 L_s C_p + j\omega R C_p} \tag{11.17}$$

where:

R = nominal value of the resistor in [Ω]
L_s = parasitic series inductance in [H]
C_p = parasitic parallel (shunt) capacitance in [F]
$\omega = 2\pi f$ = signal frequency in [rad/s]

Besides the parasitic elements of L_s [H] and C_p [F], the frequency response of a resistor also depends on the resistor's nominal value itself. Figure 11.10a shows the simplified frequency response for resistors with small nominal values and Fig. 11.10b for large resistor values.

Nowadays, film resistors are widely used because they are inexpensive and accurate. Figure 11.11 shows a simulation of the frequency behavior of different types of film resistors. With increasing frequency, the parasitic parallel (shunt) capacitance C_p [F] and the parasitic series inductance L_s [H] lead to nonideal frequency response. The plots in Fig. 11.11 show that for frequencies $f > 100\,\text{MHz}$, only surface-mount technology (SMT) resistors should be used (axial leaded resistors are not suitable for high-frequency applications due to the wired terminals). Above

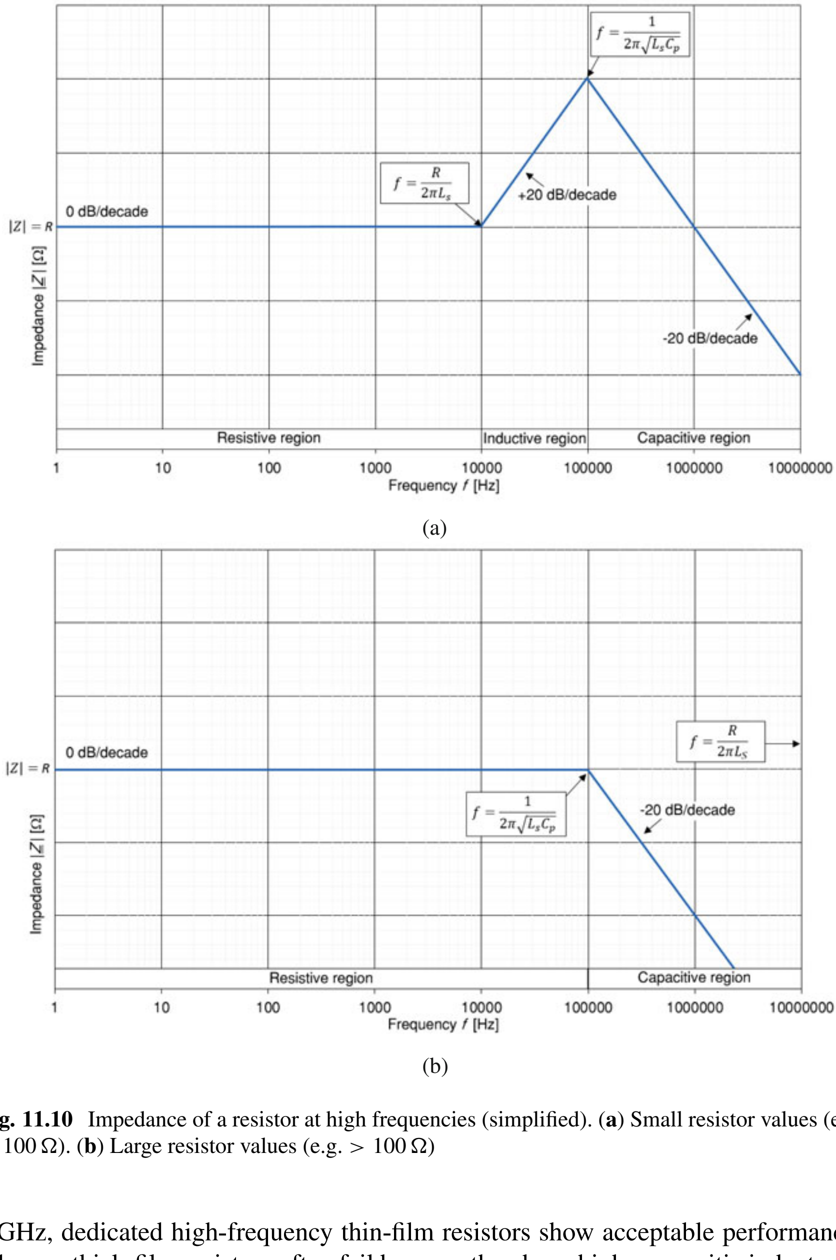


Fig. 11.10 Impedance of a resistor at high frequencies (simplified). (**a**) Small resistor values (e.g. $< 100\,\Omega$). (**b**) Large resistor values (e.g. $> 100\,\Omega$)

1 GHz, dedicated high-frequency thin-film resistors show acceptable performance, whereas thick-film resistors often fail because they have higher parasitic inductance and capacitance compared to thin-film resistors.

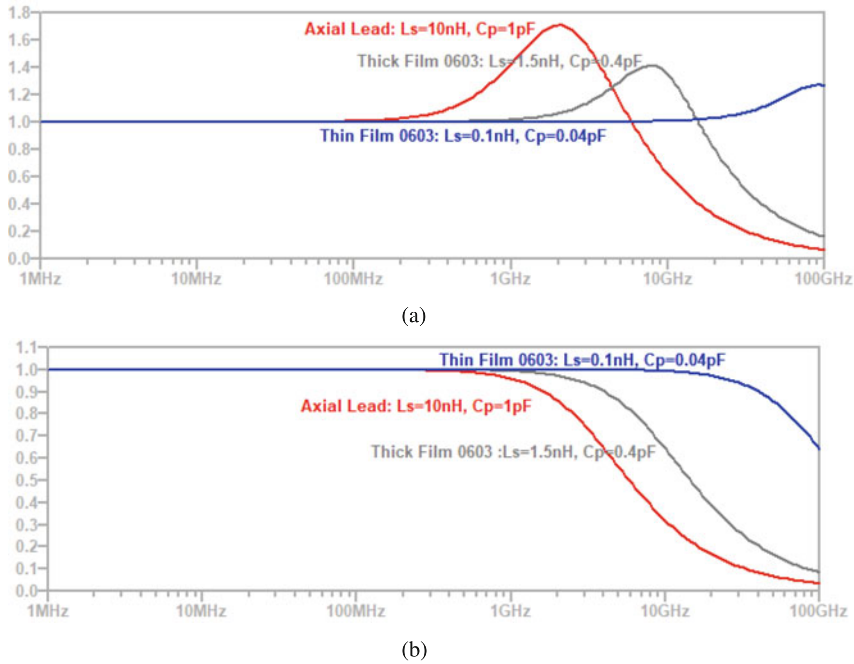


Fig. 11.11 Impedance $|Z|/R$ vs. frequency f [Hz] of different film resistors (simulation, skin effect not considered). (**a**) 50 Ω film resistors. (**b**) 1 kΩ film resistors

11.3 Capacitors

Capacitors can be categorized by its dielectric material:

- **Ceramic.** *Multilayer ceramic capacitors* (MLCCs) are the most widely used capacitors today. They have relatively low *equivalent series inductance* (ESL) and low *equivalent series resistance* (ESR). They are used up to several GHz (dielectric material C0G, NP0).
- **Electrolytic.** Electrolytic aluminum and tantalum capacitor have a high capacitance-to-volume ratio and they have relatively high ESRs. They are usually used up to 25 kHz or 100 kHz.
- **Film and Paper.** Film and paper capacitors have considerably lower ESR than electrolytic capacitors but still moderately large inductance. They are usually used up to several MHz.

Moreover, capacitors can be categorized according to their application:

- **Decoupling.** Decoupling or bypass capacitors are placed close to the power supply pins of integrated circuits (IC) to provide a stable supply voltage to the ICs. In addition, decoupling capacitors help prevent voltage drops or spikes

Table 11.1 Safety capacitor classification [15]

Subclass	Peak values of surge voltage V_p
X1	C≤1 μF: V_p = 4kV C>1 μF: V_p = 4kV/√C (C in μF)
X2	C≤1 μF: V_p = 2.5kV C>1 μF: V_p = 2.5kV/√C (C in μF)
X3	No test

(a) X-capacitor classification.

Subclass	Peak values of surge voltage V_p
Y1	8 kV
Y2	5 kV
Y3	Not pulse rated
Y4	2.5 kV

(b) Y-capacitor classification.

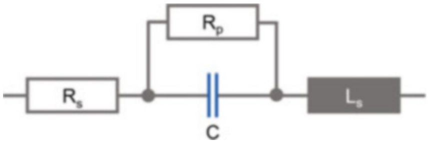


Fig. 11.12 Simplified equivalent circuit of a capacitor [4]

on the power supply line in case of a sudden change of current drawn by the integrated circuit. Usually, MLCCs with X7R or X5R dielectric are used as decoupling capacitors.

- **Low-voltage signal filters.** Signal filters with a defined cutoff frequency in the low-voltage range (< 50 V) are usually performed with stable class I MLCCs (C0G, NP0).
- **AC mains power filters.** AC mains power filters require capacitors with a sufficient safety rating (IEC 60384-14; USA: UL 1414 and UL 1283; Canada: CAN/CSA C22.2 and CAN/CSA 384-14; China: GB/T 14472) because a failure of a mains power filter capacitor could result either in fire (short circuit of an X-capacitor) or in electric shock (short circuit of a Y-capacitor). The safety classifications for X- and Y-capacitors according to IEC 60384-14 [7] are presented in Table 11.1.

The impedance $\underline{Z}$ [Ω] of an ideal capacitor is:

$$\underline{Z} = \frac{1}{j\omega C} \tag{11.18}$$

Figure 11.12 shows a simplified equivalent circuit model of a nonideal capacitor. A capacitor is not a pure capacitance C [F]. The series inductance L_s (ESL) is caused by the leads and the internal structure. The leads and internal losses also cause the series resistance R_s (ESR). The parallel resistor R_p [Ω] represents the nonideal dielectric material (leakage current) and has typically a value of 10 GΩ or more [8]. The impedance $\underline{Z}$ [Ω] of a nonideal capacitor can be written as (with neglected R_p [Ω]):

$$\underline{Z} = \frac{1 - \omega^2 L_s C + j\omega R_s C}{j\omega C} \tag{11.19}$$

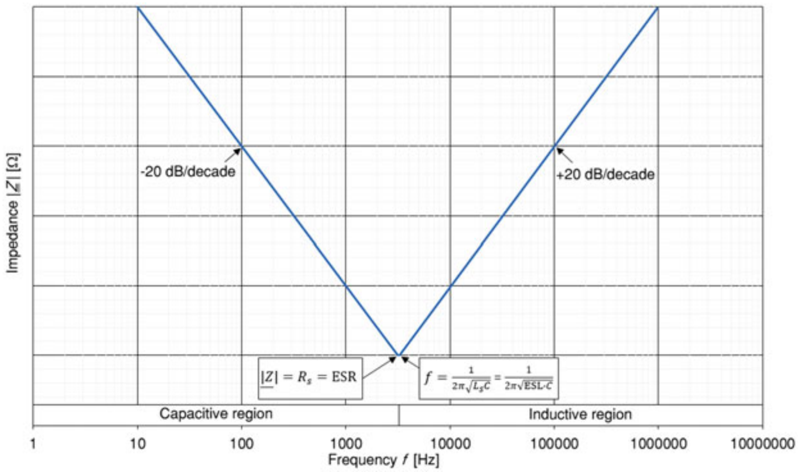


Fig. 11.13 Impedance of a capacitor at high frequencies (simplified)

where:

C = nominal capacitance in [F]
R_s = parasitic series resistance in [Ω]
L_s = parasitic series inductance in [H]
$\omega = 2\pi f$ = signal frequency in [rad/s]

Figure 11.13 shows the frequency response (impedance vs. frequency) of a real, nonideal capacitor. The capacitance dominates from DC to the serial resonant frequency f_r [Hz] of L_s [H] and C [F]. At the capacitor's resonance frequency f_r [Hz], the impedance of a capacitor is equal to R_s (ESR). However, the inductance L_s [H] dominates for frequencies higher than the resonant frequency, and the capacitor becomes an inductor. R_p [Ω] is usually neglected when it comes to high-frequency response calculations of capacitors. Note: the ESR is not simply a resistance that can be measured with an ohmmeter. The ESR represents the ohmic and dielectric losses of a capacitor.

Terms that are often used in connection with the nonideal behavior of capacitors are dissipation factor (DF), loss tangent $\tan(\delta_e)$ (Fig. 11.14), and quality factor Q, and they are defined like this:

$$\tan(\delta_e) = \frac{\mathrm{ESR}}{X_C} = \frac{1}{Q} \tag{11.20}$$

$$\mathrm{DF} = \tan(\delta_e) = \frac{1}{Q} \tag{11.21}$$

$$Q = \frac{X_C}{\mathrm{ESR}} = \frac{1}{\tan(\delta_e)} = \frac{1}{\mathrm{DF}} \tag{11.22}$$

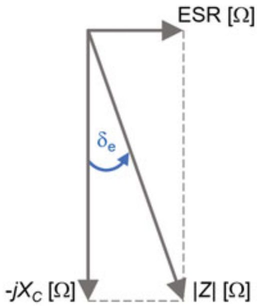


Fig. 11.14 Loss angle δ_e of a capacitor

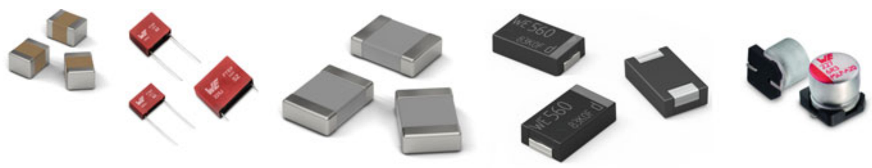


Fig. 11.15 Different form factors of capacitors. Courtesy of Würth Elektronik GmbH

where:

δ_e = loss angle of the capacitor in [rad]
$\tan(\delta_e)$ = loss tangent
ESR= equivalent series resistor of the capacitor in [Ω]
$X_C = 1/(\omega C)$ = reactance of the capacitor in [Ω]
DF= dissipation factor
Q = quality factor

The higher the dissipation factor (DF) and the loss tangent $\tan(\delta_e)$, the higher the losses of a capacitor.

Figure 11.15 presents form factors of various capacitors, and Fig. 11.16 compares the impedance behavior of C0G capacitors.

For capacitor values $<100\,\mu\text{F}$, multilayer ceramic capacitors (MLCCs) are in most cases the best option. MLCCs have excellent small ESR and ESL and are inexpensive, accurate, robust, stable long term, and available in small packages (e.g., in the SMT packages 0201, 0402). However, MLCCs have also their drawbacks which must be considered [10]:

- **Temperature derating.** The capacitance of MLCCs varies with temperature T [°C].
  - Class I. Dielectrics like C0G (NP0) and U2J are ultra-stable across a wide temperature range from $-55\,°\text{C}$ to $+125\,°\text{C}$. Table 11.2 shows the codes of class I MLCCs.
  - Class II. Dielectrics like X7R and X5R have a much higher dielectric constant ϵ_r' than class I dielectrics, but the capacitance can vary as much as $\pm 15\,\%$

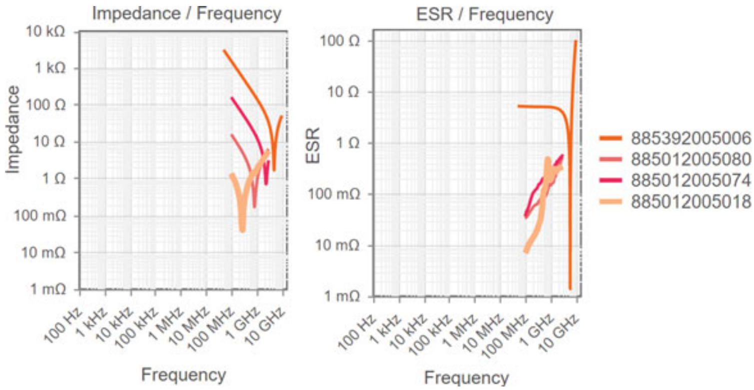


Fig. 11.16 Impedance $|\underline{Z}|$ [Ω] vs. frequency f [Hz] of C0G SMT capacitors in a 0402 case: 1 pF, 10 pF, 100 pF, 1 nF. Courtesy of Würth Elektronik GmbH

Table 11.2 Class I ceramic capacitor types. For example, C0G: ±30 ppm/K, U2J: −750 ppm/K ±120 ppm/K [10]

Letter code	Temperature coefficient α [ppm/K]	Multiplier of temperature coefficient α	Tolerance of temperature coefficient α [ppm/K]
C	0	0: −1	G: ±30
B	0.3	1: −10	H: ±60
L	0.8	2: −100	J: ±120
A	0.9	3: −1000	K: ±250
M	1.0	4: +1	L: ±500
P	1.5	6: +10	M: ±1000
R	2.2	7: +100	N: ±2500
S	3.3	8: +1000	
T	4.7		
V	5.6		
U	7.5		

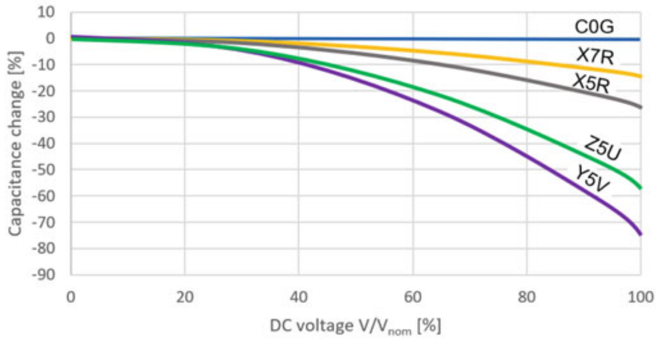
over the range of −55 °C to +85 °C (X5R) or to 125 °C. Table 11.3 shows the codes of class II and class III MLCCs.

– Class III. Dielectrics like Y5V and Z5V have the largest dielectric constants ϵ_r', but the capacitance can vary as much as +22 % to −56 % for Z5U (10 °C to +85 °C) or even +22 % to −82 % for Y5V (−30 °C to +85 °C).

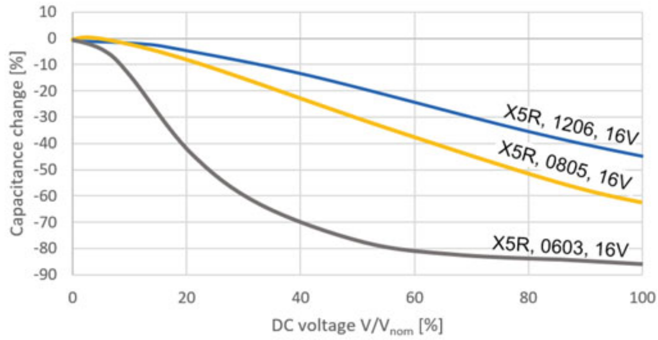
- **Voltage derating.** Class I dielectrics show no voltage derating effect (meaning, the capacitance of the C0G (NP0) or U2J capacitor does not change with the applied DC voltage). However, class II and class III show a significant drop (up to −90 %) of its rated capacitance with applied DC voltage (see Fig. 11.17a). It looks different for applied AC voltages. The capacitance of a capacitor is usually specified at 1 Vrms. Class II and class III MLCCs show an increase

Table 11.3 Class II and class III ceramic capacitor types [10]. For example, X7R: ±15 % from −55 °C to +125 °C, Z5U: 22 % to −56 % from +10 °C to +85 °C

Letter code	Lower temperature of range [°C]	Numberical code	Upper temperature of range [°C]	Letter code	Maximum capacitance change over temperature range [%]	Class
Z	+10	2	+45	A	±1.0	Class II
Y	−30	4	+65	B	±1.5	
X	−55	5	+85	C	±2.2	
		6	+105	D	±3.3	
		7	+125	E	±4.7	
		8	+150	F	±7.5	
		9	+200	P	±10	
				R	±15	
				S	±22	
				L	+15 to −40	
				T	+22 to −33	Class III
				U	+22 to −56	
				V	+22 to −82	



(a)



(b)

Fig. 11.17 DC bias voltage derating of MLCCs. (**a**) MLCC DC bias voltage derating vs. dielectric material. Exact values vary. (**b**) MLCC DC bias voltage derating vs. case size. Exact values vary

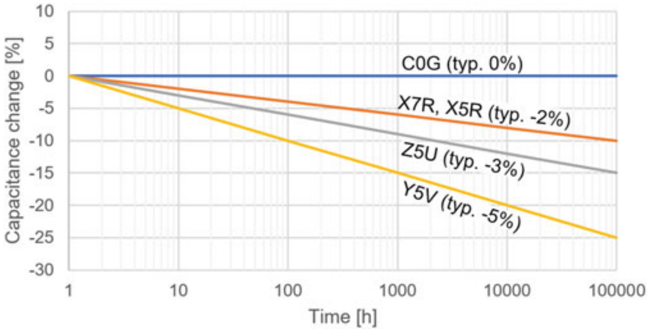


Fig. 11.18 Typical aging of MLCCs. Exact values can vary

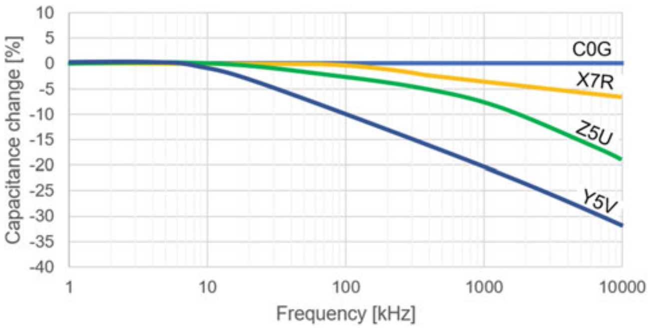


Fig. 11.19 Typical capacitance derating of MLCCs due to signal frequency. Exact values vary

of capacitance when an AC voltage within a reasonable range is applied and a decrease for small AC voltage amplitudes (e.g., $-10\,\%$ at $10\,\mathrm{mV}$). If a high enough AC voltage is applied, eventually, it will reduce capacitance just as a DC voltage. Capacitance decreases more quickly with smaller case sizes (see Fig. 11.17b).

- **Aging.** Class I dielectrics show very little to no aging effect [10], whereas class II and class III dielectrics have an aging effect (see Fig. 11.18). For example, X7R and X5R have a typical aging rate $A = 2\,\%/\mathrm{decade} - \mathrm{hours}$, Z5U $A = 3\,\%/\mathrm{decade} - \mathrm{hours}$ and Y5V $A = 5\,\%/\mathrm{decade} - \mathrm{hours}$ (a decade-hour is, e.g., from 1 h to 10 h or from 100 h to 1000 h).
- **Frequency.** Class I dielectrics do not show a frequency-dependent capacitance. On the other hand, class II and class III dielectric materials show a frequency-dependent capacitance (see Fig. 11.19).

The conclusion from the points above is that whenever a stable capacitance C [F] is needed, use a class I MLCC capacitor (e.g., NP0, C0G for filter applications). Another important point to remember is that the capacitance C [F] of class II and class III MLCCs is a function of temperature T [°C], DC voltage V [V], time (aging) t [h], and frequency f [Hz]. Nonetheless, class II and class III MLCCs have their use cases. For example, X7R capacitors are commonly used for decoupling on PCB designs because X7R capacitors provide a higher nominal capacitance C [F] than class I capacitors for the same package size.

For capacitor values $> 100\,\mu F$, tantalum, aluminum electrolyte, or polymer electrolyte capacitors are the most popular. However, we do not discuss any other capacitor types than MLCCs because the trend is clearly toward MLCCs due to their small size, high reliability, long lifetime, and low ESL at relatively low costs.

11.4 Inductors

Inductors can be categorized in:

- **Nonmagnetic core inductors.** For example, air-core inductors, which are often used for high-frequency applications because air-core inductors are free from core losses that occur in ferromagnetic cores at increasing frequency.
- **Magnetic core inductors.** For example,, ferrite-core inductors or toroidal inductors (closed-loop magnetic core inductors).

The impedance $\underline{Z}$ [Ω] of an ideal inductor is:

$$\underline{Z} = j\omega L \tag{11.23}$$

The simplified equivalent circuit model of a nonideal inductor is shown in Fig. 11.20. The ideal inductance L [H] is influenced by the winding capacitance C_p [F] and the ohmic series resistance of the winding conductor R_s [Ω]. R_p [Ω] represents the core losses and can be neglected for nonmagnetic cores (air instead of a magnetic core).

The impedance $\underline{Z}$ [Ω] (Fig. 11.21) of a nonideal inductor can be written as:

$$\underline{Z} = \frac{R_s + j\omega L}{1 - \omega^2 LC_p + j\omega R_s C_p}, \; R_p \text{ neglected} \tag{11.24}$$

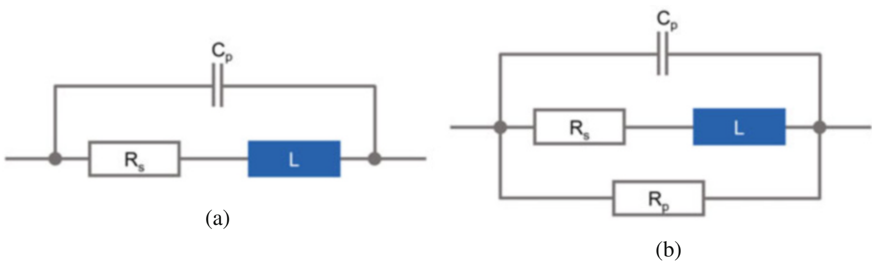


Fig. 11.20 Simplified high-frequency equivalent circuit of an inductor [4, 5]. (**a**) Air core inductor. (**b**) Magnetic core inductor

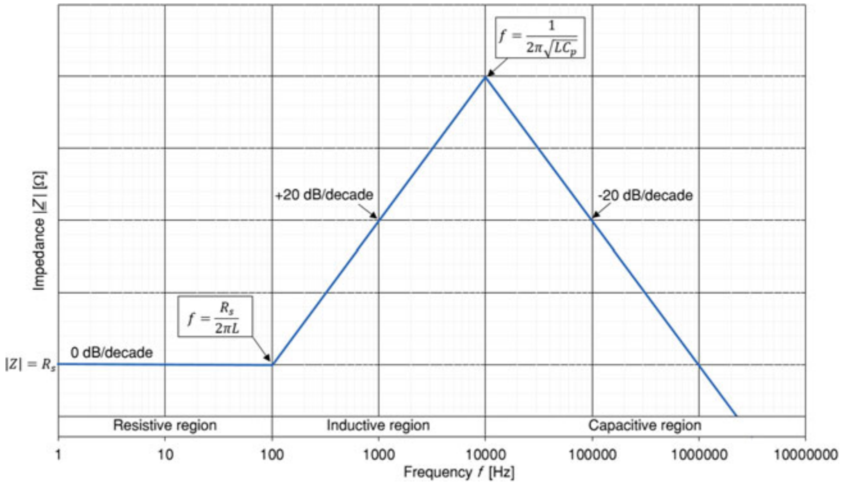


Fig. 11.21 Impedance of an inductor at high frequencies (simplified)

$$\underline{Z} = \frac{R_s + j\omega L}{1 + R_s/R_p - \omega^2 LC_p + j\omega\left(L/R_p + R_s C_p\right)}, R_p \text{ considered} \tag{11.25}$$

where:

L = nominal self-inductance in [H]
C_p = parasitic parallel (shunt) capacitance in [F]
R_s = series resistance of the windings in [Ω]
R_p = parallel resistance representing the magnetic losses in [Ω]
$\omega = 2\pi f$ = signal frequency in [rad/s]

Self-inductance L [H] is defined as the ratio of the total magnetic flux $\Phi(t)$ [Wb] to the current $i(t)$ [A], which causes that magnetic flux. Assuming a magnetic lossless material through which the magnetic flux flows ($\mu' = \mu_r'\mu_0$, $\mu'' = 0$), we can write:

$$L = \frac{\Phi(t)}{i(t)} = \frac{\vec{B}\cdot\vec{A}}{i(t)} = \frac{\left(\mu'\vec{H}\right)\cdot\vec{A}}{i(t)} \tag{11.26}$$

where:

$\Phi(t)$ = magnetic flux produced by the current $i(t)$ through inductor in [Wb]
$i(t)$ = current through the inductor in [A]
$\vec{B}$ = magnetic flux density vector in [T]
$\vec{H}$ = magnetic field vector in [A/m]
$\vec{A}$ = vector area through which the magnetic flux flows in [m^2]

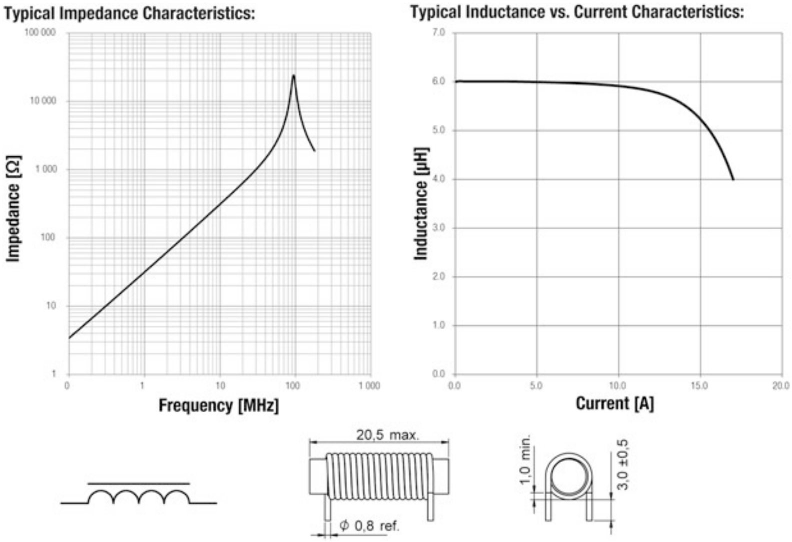


Fig. 11.22 Impedance $|\underline{Z}|$ [Ω] vs. frequency f [Hz] of the filter inductor Würth Elektronik 7847121060. Courtesy of Würth Elektronik GmbH

$\mu' = \mu'_r \mu_0$ = magnetic permeability of the material through which the magnetic field flows [H/m]

Equation 11.26 shows that the self-inductance L [H] is proportional to the core material's relative magnetic permeability μ'_r. It is important to note that μ'_r and therefore μ' and the self-inductance L [H] are a function of:

- **Current.** The higher the current I [A], the lower the $\mu_r(I)$ becomes. This means that the inductance L [H] decreases with increasing current I [A] through an inductor. Therefore, care must be taken that an inductor is used below its defined saturation current I_{sat} [A].
- **Frequency.** Typical core materials are manganese-zinc (MnZn) and nickel-zinc (NiZn). MnZn cores tend to have high initial magnetic permeability μ'_{ri}, but their $\mu'_r(f)$ deteriorate more rapidly with increasing frequency f [Hz] than that of NiZn cores. The frequency behavior of an inductor is usually presented in its datasheet.
- **Temperature.** The relative magnetic permeability $\mu'_r(T)$ of any ferromagnetic material changes with temperature T [°C]. Usually, the permeability μ'_r peaks out just before the material reaches its *Curie temperature* [5]. At the Curie temperature, the magnetic material loses all its permanent magnetic properties.

As an example for the nonideal behavior of inductors, Fig. 11.22 presents the frequency-dependent impedance $|\underline{Z}|$ [Ω] and the current-dependent inductance L [H] of a rod-inductor.

With the introduction of the complex magnetic permeability (Fig. 11.23):

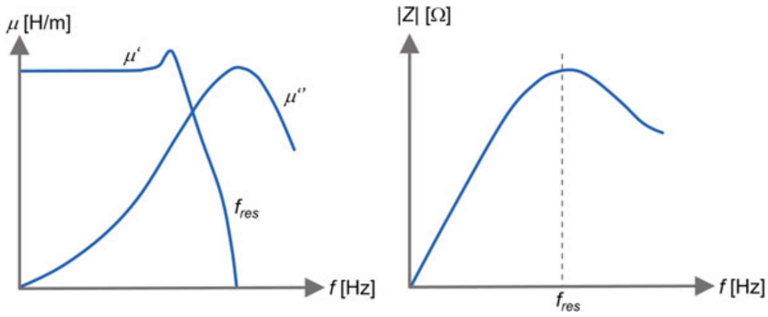


Fig. 11.23 Typical frequency response of an inductor and its complex magnetic permeability $\underline{\mu} = \mu' - j\mu''$ and its impedance $|\underline{Z}|$ [5]

Fig. 11.24 Equivalent circuit of an inductor when considering magnetic losses only and ignoring ohmic losses and winding capacitance

$$\underline{\mu} = \mu' - j\mu'' \tag{11.27}$$

where:

$\mu' = \mu_r'\mu_0$ = real part of the complex permeability in [H/m]
$\mu'' = \mu_r''\mu_0$ = imaginary part of the complex permeability in [H/m]
$j = \sqrt{-1}$

we can separate the ideal self-inductance L [H] and the magnetic losses of the core material R_{ml} [Ω]. This concept can be applied to all types of inductors and ferrites. Now we can write:

$$\underline{Z} = j\omega\underline{\mu}L_0 = j\omega L_0\left(\mu' - j\mu''\right) = R_{ml} + jX_{ind} \tag{11.28}$$

$$R_{ml} = \omega L_0\mu'', \text{ magnetic loss component} \tag{11.29}$$

$$X_{ind} = \omega L_0\mu', \text{ inductance component} \tag{11.30}$$

where:

$\underline{\mu} = \mu' - j\mu''$ = complex magnetic permeability of the inductor in [H/m]
$\omega = 2\pi f$ = signal frequency in [rad/s]
L_0 = inductance of an air inductor of the same construction and field distribution, without core material ($\underline{\mu} = \mu_0$, $\mu_r' = 1$) in [H]
R_{ml} = magnetic losses modeled as series resistor to the inductance in [Ω]
X_{ind} = inductive reactance in [Ω]

Figure 11.24 shows the equivalent circuit of an inductor when ohmic losses R_s [Ω] and capacitive parasitics C_p [F] are ignored. The loss tangent $\tan(\delta_m)$ of this

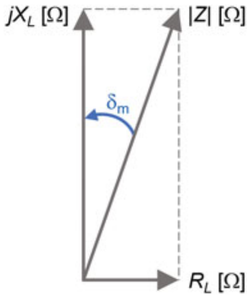


Fig. 11.25 Loss angle δ_m of an inductor

setup is given as:

$$\tan\left(\delta_m\right) = \frac{R_{ml}(f)}{X_{ind}(f)} = \frac{\mu''(f)}{\mu'(f)} \tag{11.31}$$

where:

$R_{ml}(f)$ = magnetic losses modeled as series resistor to the inductance in [Ω]
$X_{ind}(f)$ = inductive reactance in [Ω]

Figure 11.23 and Eq. 11.31 illustrate that the inductance L [H] is frequency-dependent. Thus, the loss tangent $\tan\left(\delta_m\right)$ (Fig. 11.25) is also frequency-dependent.

When considering all losses, the quality factor Q of an inductor is defined as:

$$Q = \frac{X_L}{R_L} = \frac{\omega L}{R_L} = \frac{1}{\tan\left(\delta_m\right)} \tag{11.32}$$

where:

X_L = reactance of the inductor (considering also parasitic elements) in [Ω]
R_L = ohmic and magnetic losses of the inductor in [Ω]
$\omega = 2\pi f$ = signal frequency in [rad/s]
$\tan\left(\delta_m\right)$ = loss tangent for loss angle δ_m

An ideal inductor would have $Q = \infty$ (lossless). However, for EMC applications, a high Q-factor is not always desired in the first place (except when acting as a filter) as this means that the inductor has low losses and a high self-inductance L [H]. In most cases, EMC applications require inductors with high losses over a wide frequency range and a Q-factor typically $Q < 3$ in the frequency range of interest [5]. This leads us directly to the next Sect. 11.5 about ferrite beads.

11.5 Ferrites Beads

Ferrite beads are magnetic components that play a crucial role in suppressing high-frequency noise and preventing electronic units from undesired radiation. Simply described, ferrite beads are electrical conductors surrounded by a magnetic material (ferrite). Ferrites are ceramic ferrite magnetic materials made from iron oxide (Fe_2O_3) and oxides of other metals. There are no eddy-current core losses (like known from iron cores) because the ferrites are made of ceramics and are electrically nonconductive. Therefore, ferrite materials can be used to provide selective attenuation of high-frequency signals.

Inductors and ferrite beads are both inductive components. However, there are some differences:

- **Inductors.** Inductors consist of a coiled electrical conductor around a magnetic or nonmagnetic core material. They typically have a much higher-quality factor Q than ferrite beads and lower losses. Typical application: filters, voltage converters.
- **Ferrite beads.** Ferrite beads consist of an electrical conductor surrounded by a ferrite material. They typically have a lower-quality factor Q than inductors and higher losses. Typical application: noise suppression (converting noise signals into heat).

The behavior of ferrite beads is often described as frequency-dependent resistance because, in the frequency range of interest, the impedance $\underline{Z}$ [Ω] of ferrite beads is dominated by its resistive part. Different types of ferrite beads can be found on the market. The two most common form factors are:

- **Cable mount ferrite beads.** Cable mount ferrite beads are wrapped around a cable, wire, or a group of conductors. They can be flexibly installed and are usually used to lower radiated emissions by suppressing common-mode currents.
- **Chip ferrite beads.** Chip ferrite beads are mounted on a PCB as SMD parts. Chip ferrite beads are compact, inexpensive, and often used to lower radiated emissions or increase immunity to radiated disturbances.

11.5.1 Cable Mount Ferrite Beads

With *cable mount ferrite beads*, the ferrite material is placed around a cable, a wire, or a group of conductors. Figure 11.26 shows typical examples of cable mount ferrite beads. In order to achieve optimal attenuation, the air gap between the cable and the surrounding ferrite bead should be minimal.

Figure 11.27 shows the simplified high-frequency equivalent circuit of a cable mount ferrite bead. The self-inductance L [H] and the magnetic losses depend on the signal frequency f [Hz]. The losses are modeled as resistance R [Ω]. The bead material can be characterized by the complex magnetic permeability [13]:

Fig. 11.26 Cable mount ferrite beads. Courtesy of Würth Elektronik GmbH

Fig. 11.27 Simplified high-frequency equivalent circuit of a cable mount ferrite bead [13]

$$\underline{\mu}(f) = \mu'(f) - j\mu''(f) \tag{11.33}$$

where:

$\mu' = \mu_r'\mu_0$ = real part of the complex permeability of the bead material (refers to the stored magnetic energy) in [H/m]
$\mu'' = \mu_r''\mu_0$ = imaginary part of the complex permeability of the bead material (refers to the magnetic losses in the bead material) in [H/m]
$j = \sqrt{-1}$

Figure 11.23 shows that $\mu'(f)$ and $\mu''(f)$ are both shown as functions of frequency f [Hz]. The complex impedance $\underline{Z}$ [Ω] of a cable mount ferrite bead can be written as [13]:

$$\underline{Z} = j\omega L_{bead} = j\omega\underline{\mu}K = j\omega\left(\mu' - j\mu''\right)K \tag{11.34}$$

$$= \underbrace{\omega\mu_0\mu_r''(f)K}_{R(f)} + j\,\underbrace{\omega\mu_0\mu_r'(f)K}_{L(f)} \tag{11.35}$$

where:

$\omega = 2\pi f$ = signal frequency in [rad/s]
$\mu'' = \mu_r''\mu_0$ = imaginary part of the complex permeability in [H/m]
$\mu' = \mu_r'\mu_0$ = real part of the complex permeability in [H/m]
K = some constant depending on the bead dimensions in [m]

Typical cable mount ferrite beads can be expected to give impedances of order 100 Ω for frequencies >100 MHz. Figure 11.28 presents the impedance $|\underline{Z}|$ [Ω] of a snap ferrite as a function of signal frequency f [Hz]. The impedance of the ferrite bead depends on:

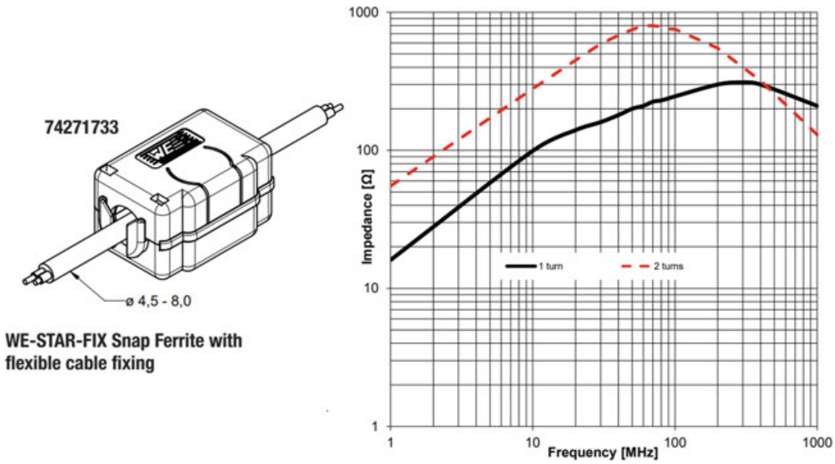


Fig. 11.28 Impedance $|\underline{Z}|$ [Ω] vs. frequency f [Hz] of the snap ferrite bead Würth Elektronik 74271733. Courtesy of Würth Elektronik GmbH

- **Frequency.** Figure 11.28 shows a typical frequency response of a snap ferrite.
- **Temperature.** The impedance $|\underline{Z}|$ [Ω] of ferrite beads reduces with increasing temperature T [°C]. Above the Curie temperature (≈120 ... 220 °C), the magnetic permeability μ_r of the ferrite becomes equal 1 (paramagnetic). The ferrite regains its previous permeability when the temperature gets back below the Curie temperature (the effect is reversible).
- **DC bias current.** The impedance $|\underline{Z}|$ [Ω] of ferrite beads reduces with increasing DC bias current I [A] because the ferrite bead core material moves toward saturation, causing a drop in inductance L [H]. The larger the volume of a ferrite core, the higher the DC bias currents can be without causing much impedance loss.

11.5.2 PCB Mount Ferrite Beads

Figure 11.29a shows some common form factors of SMT *chip ferrite beads*, whereas Fig. 11.29b shows the inner structure. Chip ferrite beads consist of a conductor surrounded by a ferromagnetic material (typical NiZn).

The simplified high-frequency equivalent circuit of a chip ferrite bead is shown in Fig. 11.30. R_{DC} [Ω] represents the DC series resistance, L [H] represents the inductance, C_p [F] represents the parasitic capacitance, and R_{AC} [Ω] represents the intended losses at high frequencies. Figure 11.31 shows a typical frequency response of a PCB mount ferrite bead:

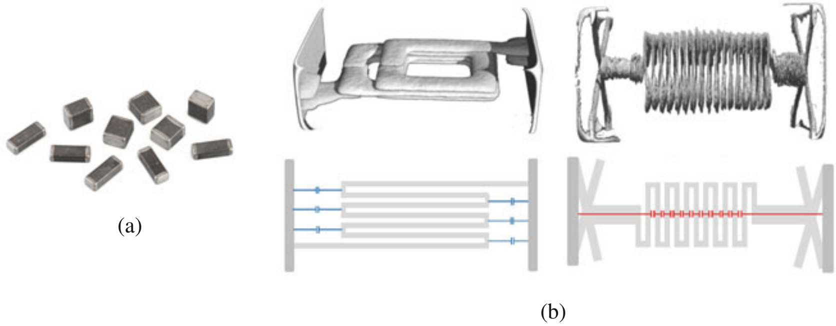
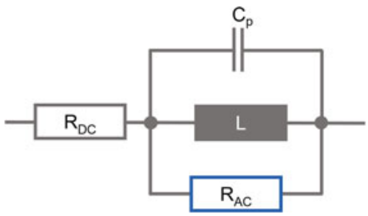


Fig. 11.29 SMT chip ferrite beads. Courtesy of Würth Elektronik GmbH. (**a**) Different form factors of SMT chip ferrite beads. (**b**) Inner structure of chip ferrite beads. Left: WE-CBF. Right: WE-CBF HF (optimized for high-frequency applications due to low C_p)

Fig. 11.30 High-frequency equivalent circuit of a SMT ferrite chip bead



- $<$ **f$_r$.** From DC until the resonance frequency f_r [Hz], the behavior of chip ferrite beads is inductive.
- $\approx$ **f$_r$.** Around f_r [Hz], the chip ferrite bead is primarily resistive ($|\underline{Z}| \approx R$). The intended use of a ferrite bead for EMI applications is that the component works in this resistive area, where the noise suppression is most effective (converting noise to heat).
- $>$ **f$_r$**. For frequencies higher than f_r [Hz], the ferrite bead impedance behavior is capacitive due to the parasitic C_p [F].

The nonideal behavior of PCB mount ferrite beads comprises the following point:

- **Frequency dependence.** As shown in Fig. 11.31, the chip ferrite bead behaves capacitive for frequencies $> f_r$, and the noise suppression for this frequency range is reduced.
- **Temperature influence.** The impedance $|\underline{Z}|$ [Ω] of ferrite beads reduces with increasing temperature T [°C] (see Fig. 11.32b). Above the Curie temperature ($\approx 120 \ldots 220\,°\text{C}$), the magnetic permeability μ_r of the ferrite becomes equal 1 (paramagnetic). The ferrite regains its previous permeability when the temperature gets back below the Curie temperature (the effect is reversible).
- **DC bias current influence.** The impedance $|\underline{Z}|$ [Ω] of ferrite beads reduces with increasing DC bias current I [A] (see Fig. 11.32a), because the ferrite bead core material moves toward saturation, causing a drop in inductance L [H].

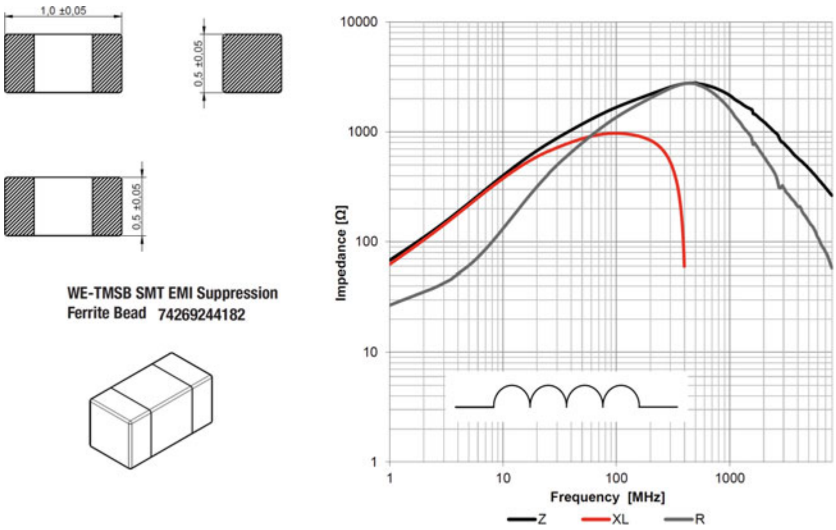


Fig. 11.31 Impedance $|\underline{Z}|$ [Ω] vs. frequency f [Hz] of the EMI suppression chip ferrite bead Würth Elektronik 74269244182. Courtesy of Würth Elektronik GmbH

11.6 Common-Mode Chokes

A *common-mode choke* consists of two coils wound around a common ferromagnetic core (sometimes even more than two coils). The differential signal current flows through one coil and back through the second one (see Fig. 11.34b). Common-mode chokes in EMC are primarily used to block common-mode currents and, therefore, to eliminate unintended radiated electromagnetic emissions.

Figure 11.33d presents the high-frequency equivalent circuit of a common-mode choke, where R_s [Ω] represents the ohmic losses, L [H] represents the self-inductance of each coil, M [H] represents the mutual inductance of the coils, R_p [Ω] represents the magnetic core losses, C_p [F] represents the parasitic capacitance between the turns of the coils, and C_{12} [F] represents the parasitic capacitance between the two coils.

When ignoring the parasitic capacitances and the ohmic and magnetic losses, the coil impedances of an ideal common-mode choke can be written as [13]:

$$\underline{Z}_1 = \frac{\underline{V}_1}{\underline{I}_1} = \frac{j\omega L_1 \underline{I}_1 + j\omega M_{12} \underline{I}_2}{\underline{I}_1} \tag{11.36}$$

$$\underline{Z}_2 = \frac{\underline{V}_2}{\underline{I}_2} = \frac{j\omega L_2 \underline{I}_2 + j\omega M_{21} \underline{I}_1}{\underline{I}_2} \tag{11.37}$$

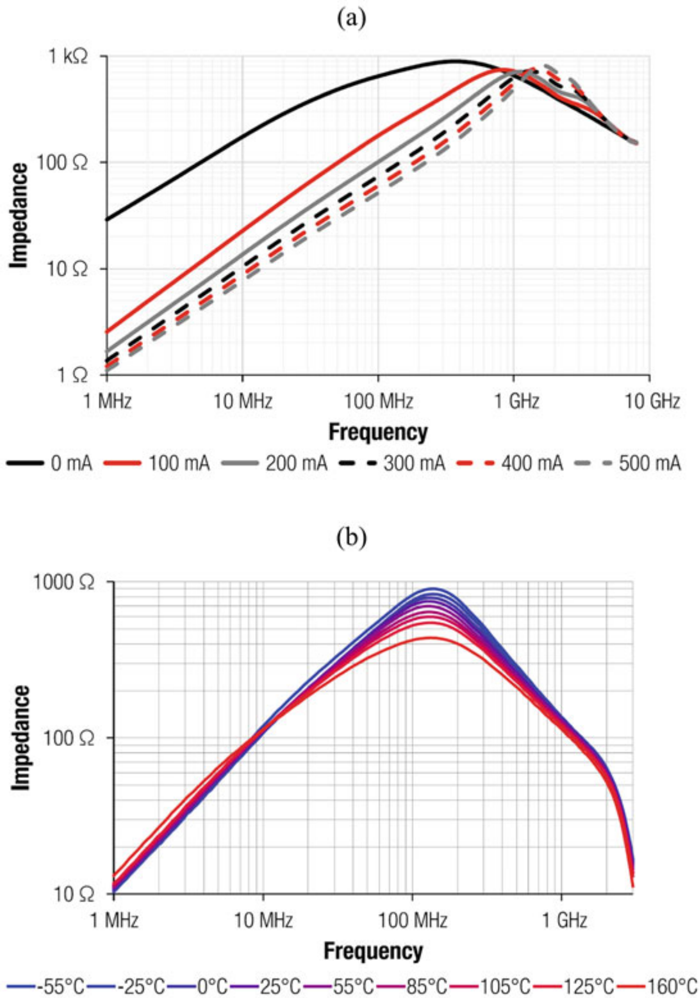


Fig. 11.32 Nonideal behavior of ferrite beads. Courtesy of Würth Elektronik GmbH [1]. (**a**) Impedance $|\underline{Z}|$ [Ω] vs. DC bias current I [A] of Würth WE-CBF 742861160. (**b**) Impedance $|\underline{Z}|$ [Ω] vs. temperature T [°C] of Würth WE-CBF 742792040

where:

$\underline{Z}_1$ = impedance of common-mode choke coil 1 in [Ω]
$\underline{Z}_2$ = impedance of common-mode choke coil 2 in [Ω]
$\underline{V}_1$ = voltage over coil 1 in [V]
$\underline{V}_2$ = voltage over coil 2 in [V]
$\underline{I}_1$ = current through coil 1 in [A]
$\underline{I}_2$ = current through coil 2 in [A]

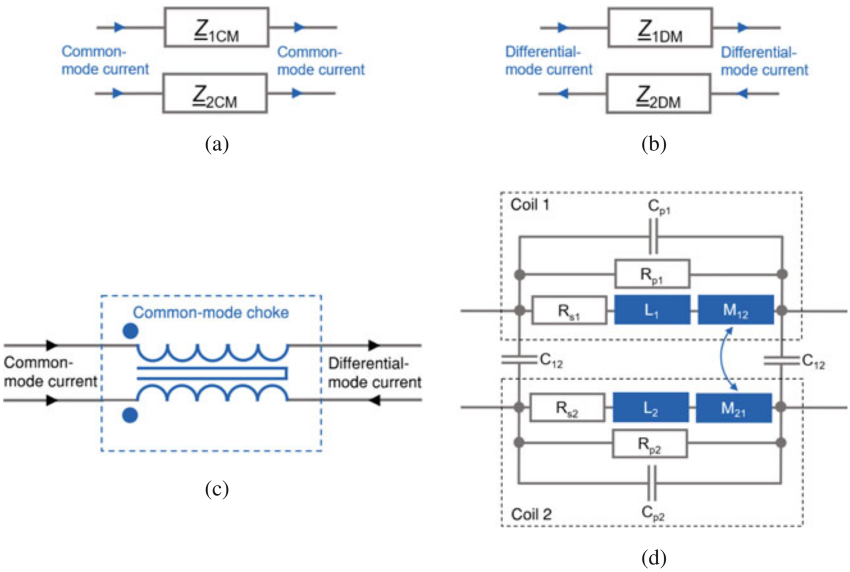


Fig. 11.33 Common-mode choke. (**a**) Common-mode impedances of a common-mode choke. (**b**) Differential-mode impedances of a common-mode choke. (**c**) Common-mode choke symbol. (**d**) High-frequency equivalent circuit of a common-mode choke

L_1 = self-inductance of coil 1 in [H]
L_2 = self-inductance of coil 2 in [H]
M_{12} = mutual inductance at coil 1 due to coupling with coil 2 in [H]
M_{21} = mutual inductance at coil 2 due to coupling with coil 1 in [H]
$\omega = 2\pi f$ = signal frequency in [rad/s]

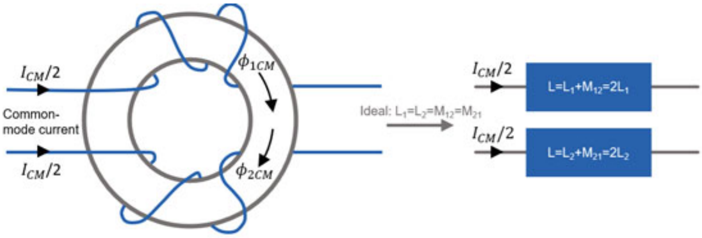
If we now assume that the common-mode current I_{CM} [A] splits up equally through both coils of the common-mode choke ($I_{1CM} = I_{2CM} = I_{CM}/2$) and that the self-inductance of the coils is equal to the mutual inductance ($L_1 = L_2 = M_{12} = M_{21}$; see Fig. 11.34a), the common-mode impedance of each coil is:

$$\underline{Z}_{1CM} = j\omega(L + M) = j\omega 2L \tag{11.38}$$

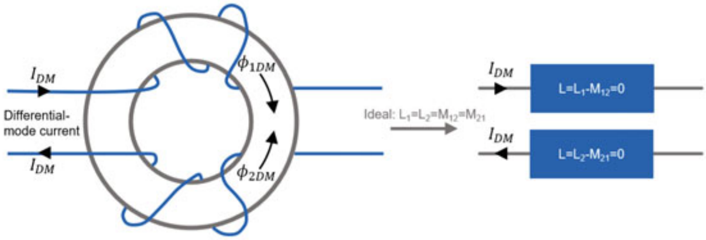
$$\underline{Z}_{2CM} = j\omega(L + M) = j\omega 2L \tag{11.39}$$

where:

$\underline{Z}_{1CM}$ = ideal common-mode impedance of common-mode choke coil 1 in [Ω]
$\underline{Z}_{2CM}$ = ideal common-mode impedance of common-mode choke coil 2 in [Ω]
L = self-inductance of a single coil of the common-mode choke in [H]
M = mutual inductance between the coils of the common-mode choke in [H]
$\omega = 2\pi f$ = signal frequency in [rad/s]



(a)



(b)

Fig. 11.34 Common-mode choke models for common-mode currents I_{CM} vs. differential-mode currents I_{DM} [13]. Parasitic capacitances (C_p, C_{12}) and losses (R_s R_p) were ignored. (**a**) I_{CM} = common-mode current. (**b**) I_{DM} = differential-mode current

Equations 11.38 and 11.39 show why common-mode chokes have a high impedance for common-mode currents. On the other hand, an ideal common-mode choke ($L_1 = L_2 = M_{12} = M_{21}$; see Fig. 11.34b) would not influence the differential-mode current at all:

$$\underline{Z}_{1DM} = j\omega(L - M) = 0 \tag{11.40}$$

$$\underline{Z}_{2DM} = j\omega(L - M) = 0 \tag{11.41}$$

where:

$\underline{Z}_{1DM}$ = ideal impedance of common-mode choke coil 1 for differential-mode currents in [Ω]
$\underline{Z}_{2DM}$ = ideal impedance of common-mode choke coil 2 for differential-mode currents in [Ω]
L = self-inductance of a single coil of the common-mode choke in [H]
M = mutual inductance between the coils of the common-mode choke in [H]
$\omega = 2\pi f$ = signal frequency in [rad/s]

Figure 11.35 shows the different form factors of common-mode chokes, and Fig. 11.36 presents a typical frequency response of a common-mode choke.

Common-mode chokes do influence not only the common-mode signals but also the differential-mode signals (see Fig. 11.36).

The intentional behavior of common-mode chokes can be summarized like this:

- **Common-Mode current impedance.** The magnetic fluxes Φ_{1CM} [Wb] and Φ_{2CM} [Wb] through the core caused by the common-mode current I_{CM} [A] through the two coils are added up (see Fig. 11.34a). Ideally, this leads to an inductance L_{CM} [H] that is twice the common-mode choke coil inductance $L_{CM} = 2L_1 = 2L_2$ and thus leads to high attenuation of common-mode currents.
- **Differential-mode current impedance.** The magnetic fluxes Φ_{1DM} [Wb] and Φ_{2DM} [Wb] through the core caused by the differential-mode current I_{DM} [A] are subtracted (ideally to zero; see Fig. 11.34b) and do not lead to any saturation of the magnetic core material. This means that even large differential-mode currents do not lead to a saturation of common-mode chokes and thus do not affect the performance of the common-mode noise filtering.

Fig. 11.35 Different form factors of common-mode chokes as mains power line filters or low-voltage signal interfaces. Courtesy of Würth Elektronik GmbH

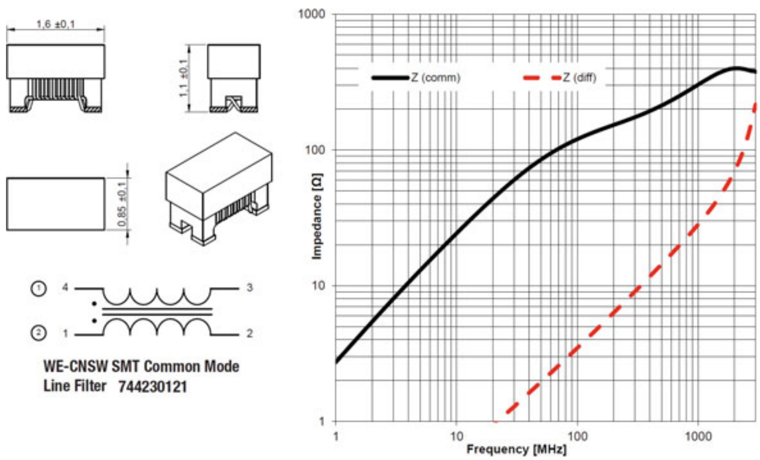


Fig. 11.36 Impedance $|\underline{Z}|$ [Ω] vs. frequency f [Hz] of the common-mode line filter 744230121 from Würth. Courtesy of Würth Elektronik GmbH

The nonideal (and therefore unintended) behavior of common-mode chokes is summarized here:

- **Frequency dependence.** As Fig. 11.36 shows, the common-mode choke influences differential signals at higher frequencies and the common-mode impedance drops. Therefore, it is necessary to check the datasheet if the common-mode choke does not influence the differential-mode signal (which is the useful signal and should therefore not be attenuated or distorted).
- **Temperature influence.** The impedance $|\underline{Z}_{CM}|$ [Ω] of a common-mode choke reduces with increasing temperature T [°C] until it suddenly drops when the Curie temperature of the magnetic core material is reached. The temperature of common-mode chokes should usually not exceed 125 °C or 150 °C.
- **Amperage of the differential current.** Even though the magnetic flux of the differential current ideally cancels out when flowing through the common-mode choke, it has to be checked that the coils can carry the high differential currents from a thermal standpoint (heating due to ohmic losses).

11.7 Baluns

A *balun* (a contraction of balanced-unbalanced) is a component that is used to convert between a balanced signal and an unbalanced signal. A balun is a bidirectional component and can be used either way (from balanced to unbalanced and vice versa). There are two main modes of baluns:

- **Voltage balun.** The operation of a balun in voltage mode is shown in Fig. 11.37a. Voltage mode is suitable for impedance matching applications [11] and applications that require galvanic isolation.
- **Current balun.** The current mode operation of a balun is shown in Fig. 11.37c. In current mode operation, it adds impedance to unwanted common-mode current and helps to eliminate them effectively [11]).

In addition to a distinction between voltage and current baluns, there is also a distinction between isolation transformers (Fig. 11.37a) and autotransformers (Fig. 11.37b). Typical balun applications are:

- **RF transceiver to antenna interface.** A balun can act as an interface converter between a single-ended RF transceiver and a balanced antenna.
- **Unbalanced to balanced interface.** A balun can act as an interface between an unbalanced (microstrip line, single-ended signal) and a balanced transmission line (twisted-pair, differential signal) and vice versa (Fig. 11.38).
- **Impedance matching.** A balun can also be used to transform between different system impedances. In other words, a balun is an impedance matching component.

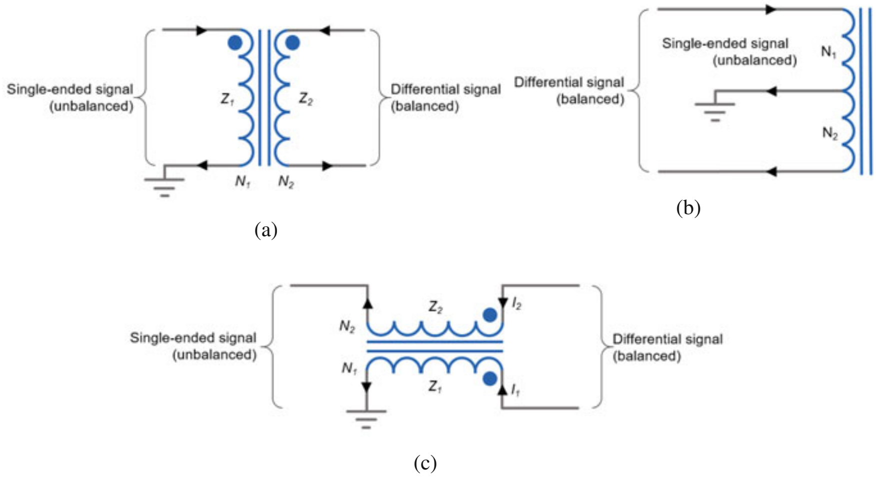
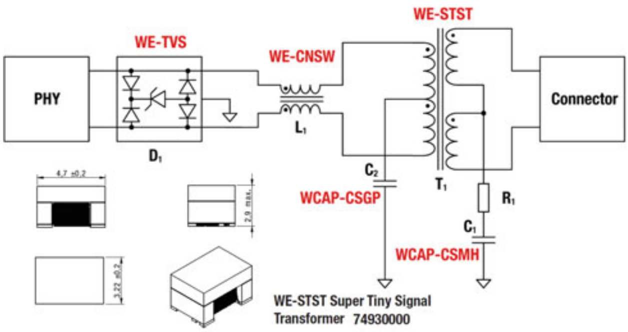


Fig. 11.37 Baluns. (**a**) Balun in voltage mode. (**b**) Autotransformer baluns are electrically non-isolating and operate with only one winding. (**c**) Balun in current mode



Properties		Test conditions	Value	Unit	Tol.
Turns Ratio	n	(1-2):(6-7), 3-4 short	1:1		±1%
DC Resistance	R_{DC}	100 kHz/ 100 mV	4.5	Ω	max.
Inductance	L	100 kHz/ 100 mV	350	µH	min.
Insertion Loss	IL	1-100 MHz	-1	dB	max.
Return Loss	RL	1-30 MHz @ 100 Ω	-18	dB	min.
Return Loss	RL	30-60 MHz @ 100 Ω	-14	dB	min.
Return Loss	RL	60-80 MHz @ 100 Ω	-12	dB	min.
Return Loss	RL	80-100 MHz @ 100 Ω	-10	dB	min.
Differential to Common Mode Rejection Ratio	DCMR	1-60 MHz	-33	dB	min.
Differential to Common Mode Rejection Ratio	DCMR	60-100 MHz	-33	dB	min.
Insulation Test Voltage	V_T	1 min.	1500	V (RMS)	min.

Fig. 11.38 10BASE-T1 and 100BASE-T1 Single Pair Ethernet (SPE) interface with galvanic isolation of 1.5 kV and balun WE-STST 74930000 from Würth. Courtesy of Würth Elektronik GmbH

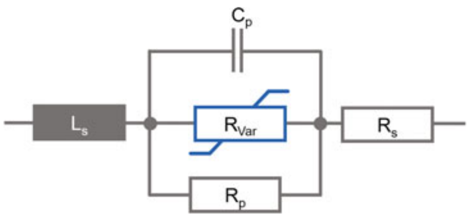


Fig. 11.39 Equivalent circuit of a varistor [5]

The applications and use cases for baluns are very broad. Therefore, it is difficult to give a conclusive set of properties to watch out for. The first thing to decide is if a voltage or a current balun or an autotransformer balun with only one coil is needed. If you are unsure which type of balun you need, you could look for an application note that describes your application best and start with the solution described there.

11.8 Clamping Devices

Voltage clamping devices —like varistors (Sect. 11.8.1) and TVS diodes (Sect. 11.8.2)—commonly block the current flow up to a specified voltage. Once the voltage reaches that clamping voltage, the voltage across the voltage clamping device does not increase anymore. In other words, the voltage is *clamped*. Voltage clamping devices are commonly used to protect circuits from transient pulses like:

- **ESD.** IEC 61000-4-2, 1/60 nsec-pulses of 1 nsec rise-time by 60 nsec duration.
- **Bursts (EFTs).** IEC 61000-4-4, 5/50 nsec-pulses of 5 nsec rise-time by 50 nsec duration.
- **Surges.** IEC 61000-4-5, 8/20 µsec-pulses of 8 µsec rise-time by 20 µsec duration.

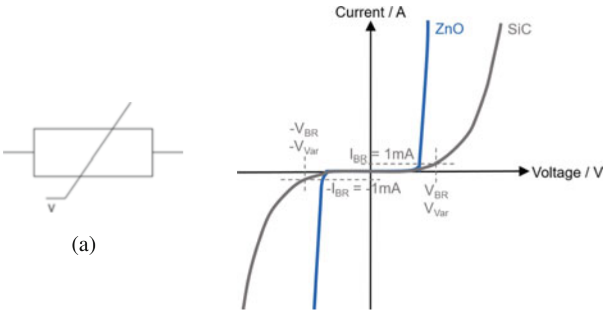
Figure 11.46 shows how a voltage clamping device limits the transient voltage across its pins.

11.8.1 Varistors

Varistors are voltage-dependent resistors (VDR), and they are typically used as overvoltage protection devices against bursts (EFTs) and surges, which arrive at the victim via the public mains supply or the DC supply port. Varistors belong to the category of clamping devices.

The high-frequency equivalent circuit for varistors is presented in Fig. 11.39. The series inductance L_s [H] represents the varistor's terminal inductance (approximately 1 nH/mm), the series resistor R_s [Ω] represents the varistor's terminal resistance (a few [mΩ]), the parallel capacitance C_p [F] represents the intergranular capacitance (10 pF up to 10 nF), the parallel resistor R_p [Ω] represents the

Fig. 11.40 Different form factors of varistors. Courtesy of Würth Elektronik GmbH



(a)

(b)

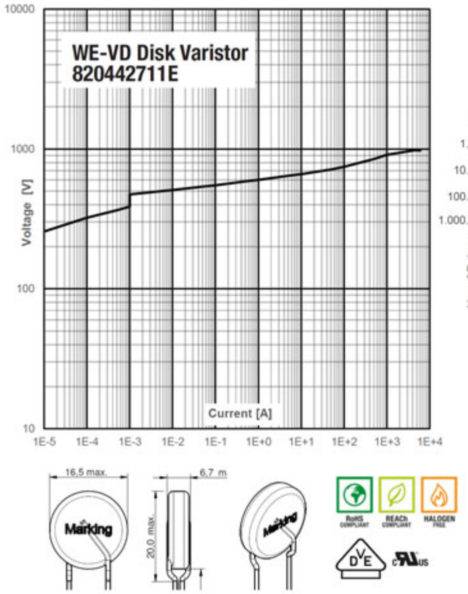
Fig. 11.41 Varistor (VDR). (**a**) Schematic symbol of varistors. The letter U is often used instead of the letter V. (**b**) Current I [A] vs. voltage V [V] of ZnO and SiC varistors. ZnO VDRs belong to the category of metal oxide varistors (MOV) and are the most common.

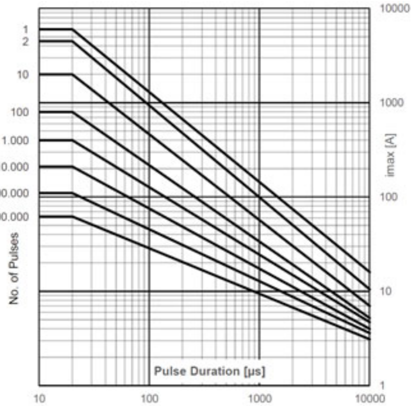
intergranular resistance (several [MΩ]), and finally R_{Var} [Ω] represents the ideal varistor ($0\,\Omega$ to $\infty\Omega$) (Fig. 11.40).

Figure 11.41b presents a varistor's nonlinear and bidirectional current-voltage characteristic, and Fig. 11.42 presents extracted data from a varistor's datasheet. At the breakdown voltage V_{BR} [V] or varistor voltage V_{Var} [V], the varistor changes its state from insulating to conducting. Usually, the breakdown voltage is specified for 1 mA DC current. Ideally, a VDR would block every current as long as the voltage across the VDR is smaller than the breakdown voltage V_{BR} [V], and as soon it is larger than V_{BR} [V], the voltage would not increase and an infinite amount of current could flow through the VDR.

Unfortunately, we do not live in an ideal world, and VDRs have some points to be considered when using them in a design:

- **Operating voltage.** Up to this voltage, the varistor does not conduct. The operating voltage in the datasheet specifies the maximum operating voltage (consider power supply tolerances when calculating the maximum operating voltage).
- **Breakdown voltage.** The breakdown voltage V_{BR} [V] specifies typically the voltage where the DC current through the varistor is 1 mA. When a voltage is applied, which exceeds V_{BR} [V], the junctions experience an avalanche breakdown, and a large current can flow.
- **Energy absorption.** Depending on the case size, the energy absorption varies from 0.01 J (SMT 0402) up to 10 kJ (chassis mount). Most varistors can absorb 5/50 nsec-burst-pulses and/or 8/20 µsec-surge pulses (take care which immunity





Properties		Test conditions	Value	Unit	Tol.
AC Operating Voltage	V_{RMS}		275	V	max.
DC Operating Voltage	V_{DC}		350	V	max.
Clamping Voltage	V_{Clamp}	50 A @ 8/20 µs	710	V	max.
(Reverse) Peak Pulse Current	I_{Peak}	8/20 µs	6000	A	max.
Power Dissipation	P_{Diss}		0.6	W	max.
Energy Absorption	W_{max}	10/1000 µs	155	J	max.
Nominal Discharge Current	I_n		3	kA	max.
Measured Limiting Voltage	V_{ML}		1110	V	max.
(Reverse) Breakdown Voltage	V_{BR}	1 mA	430	V	±10%
(Channel) Input Capacitance	C_{Ch}	1 kHz	450	pF	typ.

Fig. 11.42 Varistor 820442711E from Würth for 230 VAC applications. Courtesy of Würth Elektronik GmbH

level you need for your product: 1, 2, 3, or 4). Figure 11.40 shows different form factors and case sizes of VDRs.

- **Response time.** Most varistors have a response time that exceeds > 25 nsec [6], which makes them too slow for ESD protection (IEC 61000-4-2, 1/60 nsec-pulses). VDRs with a fast response time of < 1 nsec—which is fast enough to protect against ESD-pulses—usually have the downside that they cannot absorb enough energy for surge pulses.
- **Clamping voltage.** The clamping voltage V_{clamp} [V] describes the voltage across the VDR for a certain pulse (rise-time) and current. Unfortunately, V_{clamp} is usually much higher than the breakdown voltage V_{BR} [V]. Therefore, care must be taken that the circuit to be protected can withstand V_{clamp} [V] without any damage.
- **Capacitance.** The parallel capacitance C_p [F] of varistors is much higher than the capacitance of TVS diodes or gas discharge tubes. C_p [F] is usually in the range of 10 pF up to 10 nF [6] and, therefore, too high for high-speed signal interfaces. However, in the case of the protection of AC or DC power inputs, the parallel capacitance C_p is very welcome and can be used to replace filter capacitors.
- **Leakage current.** The leakage current depends on the voltage which is applied across the VDR, and it is usually in the range of 1 µA to <1 mA. The leakage

current can be extracted from the current-to-voltage characteristic curve for any given voltage (see Fig. 11.42).

- **Aging.** As shown in Fig. 11.42, VDRs show an aging effect based on the number of current pulses they experience.

11.8.2 TVS Diodes

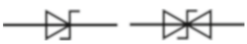
Transient-voltage-suppression diodes (TVS diodes) are typically used as overvoltage protection devices against ESD pulses, bursts (EFTs), and surges (if the current through the diode is limited). TVS diodes belong to the category of clamping devices. Figure 11.43a presents different form factors of TVS diodes.

Compared to varistors, TVS diodes have a shorter response time and a lower clamping voltage V_{clmap} [V] (for the same breakdown voltage V_{BR} [V]). On the other side, TVS diodes do endure less pulse current I_{PP} [A] compared to varistors. TVS diodes are available in two configurations:

- **Unidirectional.** Forward operation: normal silicon diode with $V_f \approx 0.7\,\text{V}$. Reverse operation: transient protection with defined breakdown voltage V_{BR} [V].
- **Bidirectional.** Both directions can be used for transient protection with defined breakdown voltage V_{BR} [V].

Figure 11.44 presents a typical current vs. voltage curve of a bidirectional TVS diode. The most essential parameters of TVS a diode are:

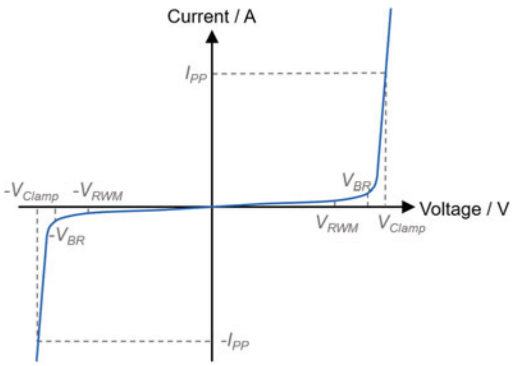
(a)

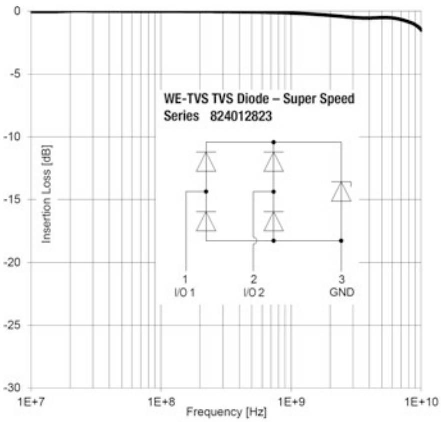


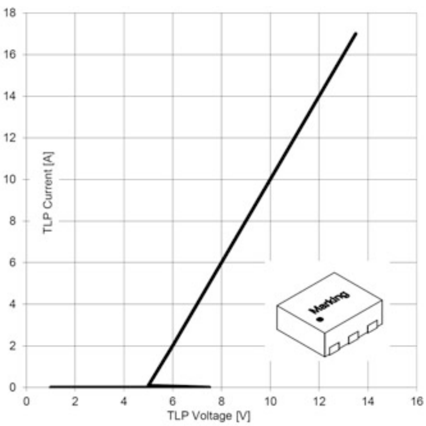
(b)

Fig. 11.43 TVS diodes. (**a**) Different form factors of TVS diodes. Courtesy of Würth Elektronik GmbH. (**b**) Left: unidirectional TVS diode. Right: bidirectional TVS diode

Fig. 11.44 Current I [A] vs. voltage V [V], a bidirectional TVS diode

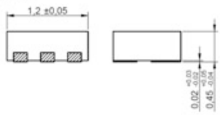






Properties		Test conditions	Value	Unit
(Reverse) Peak Pulse Current	I_{Peak}	$t_p = 8/20\mu s$	3	A
(Reverse) Peak Pulse Current	$I_{Peak\ EFT}$	$t_p = 5/50ns$	20	A
ESD Air Discharge Capability [1]	$V_{ESD\ Air}$		15	kV
ESD Contact Discharge Capability [2]	$V_{ESD\ Contact}$		8	kV
Channels	2			

[1] in positive and negative polarity
[2] in positive and negative polarity



Properties		Test conditions	Value min.	Value typ.	Value max.	Unit
Channel Operating Voltage	V_{Ch}	I/O to GND			3.3	V
(Reverse) Breakdown Voltage	V_{BR}	I/O to GND; I_{BR}=1mA	4.5			V
Channel (Reverse) Leakage Current	$I_{Ch\ Leak}$	I/O to GND $V_{I/O}=V_{Ch}$; V_{GND}= 0V			0.5	µA
Forward Voltage	V_F	GND to I/O; I_F=15mA		0.9	1.1	V
(Channel) Input Capacitance	C_{Ch}	I/O to GND $V_{I/O}$=1.65V; V_{GND}=0V; f=1MHz		0.18	0.27	pF
Channel to Channel Input Capacitance	C_{Cross}	between I/O pins $V_{I/O}$=1.65V; V_{GND}=0V; f=1MHz		0.04	0.08	pF
Channel ESD Clamping Voltage	$V_{Ch\ Clamp\ ESD}$	IEC 61000-4-2 +8kV (TLP=16A) Contact Mode; I/O to GND		13		V
Polarity	Unidirectional					

Fig. 11.45 TVS diode 824012823 from Würth for high-speed interfaces like USB 3.1 (10 GB/s) or HDMI 2.0. Courtesy of Würth Elektronik GmbH

- **Reverse standoff voltage, reverse working voltage.** Up to the reverse working voltage V_{RWM} [V], the suppressor diode does not conduct any significant amount. V_{RWM} [V] must always be greater than the maximum operating voltage (consider all the tolerances of the signal's amplitude).
- **Breakdown voltage.** At the breakdown voltage V_{BR} [V], the TVS diode starts to conduct and the diode changes to the conductive state.
- **Clamping voltage.** At the clamping voltage V_{clamp} [V], the suppressor diode carries the maximum current I_{PP} [A]. Therefore, care must be taken that the circuit to be protected can withstand V_{clamp} [V] without any damage.
- **Peak pulse current.** At the peak pulse current I_{PP} [A], the TVS diode gets the maximum energy that a short electrical pulse can convert into heat in the suppressor diode without damaging it.
- **Leakage current.** The leakage current is the current that flows in the reverse direction through the suppressor diode at a given voltage.
- **Parasitic parallel capacitance.** The parasitic parallel capacitance of the diode could lead to bad signal integrity in the case of high-speed signal interfaces.
- **Insertion loss IL.** TVS diodes designed for high-speed interfaces often specify the insertion loss IL [dB] of the TVS diodes in their datasheets (see Fig. 11.45).

11.9 Crowbar Devices

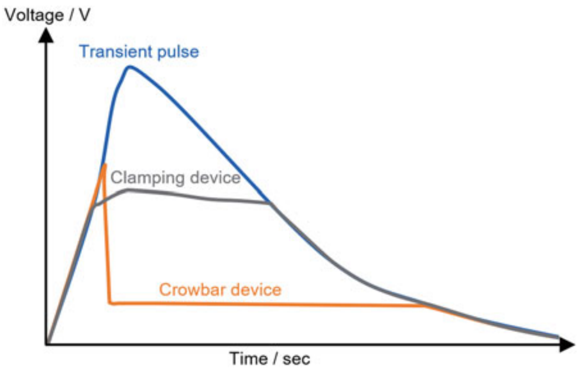
Crowbar devices commonly block the current flow up to a specific voltage. Once that voltage is exceeded, the crowbar device changes suddenly into a low impedance state, the voltage across the crowbar device gets low, and the current through the crowbar device increases and is only limited by the external circuit. After that, when the current through the crowbar device falls below a certain threshold, the crowbar device resets itself. Examples of crowbar devices are *TVS thyristors* (see Fig. 11.47b) and *gas discharge tubes* (see Fig. 11.47c). Crowbar devices are mainly used to protect circuits from high-energy transient pulses like:

- **Surges.** IEC 61000-4-5, 8/20 μsec-pulses of 8 μsec rise-time by 20 μsec duration.
- **EMP transients.** Electromagnetic pulse (EMP) transients are transients generated by EMPs that are produced, e.g., by detonations of nuclear explosive.

Compared to voltage clamping devices (see Sect. 11.8), the crowbar devices can handle much more energy, but the downside is that they do not turn on as fast as voltage clamping devices. Figure 11.46 compares the principles of voltage clamping vs. crowbar behavior (Fig. 11.47).

We do not go further into detail about crowbar devices because for many products on the market, the presented clamping devices—like varistors and TVS diodes, in combination with other passive filter components like capacitors,—are most practical and economical to use. However, gas discharge tubes are very popular for lightning protection of antennas and telecommunication products.

Fig. 11.46 Voltage pulse response of voltage clamping and crowbar devices



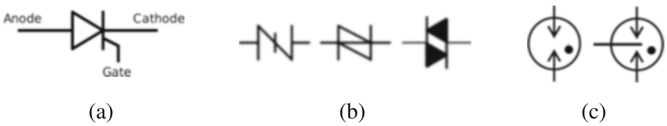


Fig. 11.47 Schematic symbols of typical crowbar devices. (**a**) Thyristor symbol. (**b**) Symbols of dedicated TVS thyristors, also called thyristor surge protection device (TSPD). (**c**) Symbols of a gas discharge tubes (GDT) with two electrodes and three electrodes

11.10 Summary

Table 11.4 summarizes the components presented in this chapter, and Table 11.5 provides a detailed comparison of transient suppression components.

Table 11.4 Summary of the high-frequency response of conductors, resistors, capacitors, inductors, ferrite beads, and common-mode chokes

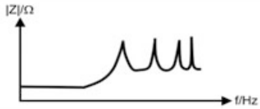
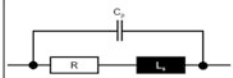
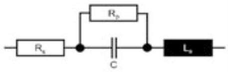
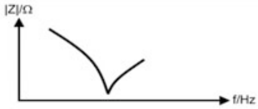
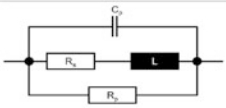
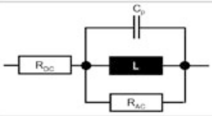
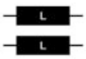
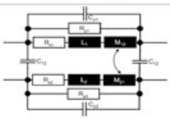
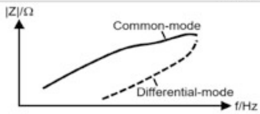
Devices	Low-Frequency (e.g. <10kHz)	High-Frequency (e.g. >10MHz)	Frequency response
Single Conductor, Single Wire			
Resistor			
Capacitor			
Inductor			
Chip Ferrite Beads			
Cable Mount Ferrite Beads			
Common-Mode Chokes			

Table 11.5 Summary of transient suppressor devices and their typical properties and applications [5, 12]

Devices	Type	Typical response time	Typical capacitance	Advantages	Disadvantages	Typical types of transients	Typical applications
TVS Diodes	Clamping	<1ns	0.1pF…10nF	- Very low cost - Low clamping voltage - Low capacitance	- Limited power handling capabilities - Typical: I_{PP}<100A	ESD burst surge	- Signal lines - DC power lines
Standard Metal Oxide Varistors (MOVs)	Clamping	<25ns	10pF…10nF	- Low cost - Large power handling capabilities (I>10kA)	- High clamping voltage - Degrade with multiple surges	Surge	- AC power lines - DC power lines
Multilayer Metal Oxide Varistors (MOVs)	Clamping	<1ns	1pF…1nF	- Higher power handling capabilities than TVS diodes - Faster switching, smaller than standard MOVs	- Lower power handling capabilities than standard MOVs - Higher clamping voltage than TVS diodes	ESD burst surge	- Signal lines
Gas Discharge Tubes	Crowbar	>1µs	0.5pF…2pF	- Large power handling capabilities (I_{PP}>10kA) - Low capacitance	- High cost - High breakdown voltage	Surge EMP-transients	- Telecom data lines - Antennas
TVS Thyristors	Crowbar	>100ns	10pF…100pF	- Typically I_{PP}>100A, up to 1kA - No degrade with multiple surges	- High cost - Higher capacitance, slower switching than TVS diodes	Surge	- Telecom data lines

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Chapter 12
Noise Coupling

The only difference between success and failure is the ability to take action.

—Alexander Graham Bell

12.1 Coupling Paths

One of the most important concepts to understand in EMC is the concept of *coupling paths*. To start off, let's see what parts are involved when electromagnetic interference (EMI) happens and why focusing on coupling paths is so important:

1. **Source.** In the real world, there are sources of unwanted electric or electromagnetic noise. For EMC immunity tests, the noise sources are well defined with the goal that these sources should be as close as possible to the real world (e.g., ESD generators, burst generators, surge generators, and antennas).
2. **Coupling path.** The noise needs a path from the source to the victim. This path is called the coupling path or coupling channel.
3. **Victim.** The victim is the receiver or receptor of the noise, which could cause interference.

Figure 12.1 shows us that if there are issues with EMI and EMC, one has the following three options to overcome these issues:

1. **Reduce the noise level of the noise source.**
    - **Emission testing.** When testing for radiated emissions of a product (meaning: the EUT is the source of noise), the noise level can be reduced by adding filters and shields to the EUT, controlling the signal transients (rise-/fall-time), or performing a complete redesign (e.g., change from single-ended to differential signaling).
    - **Immunity testing.** The applicable EMC product standard specifies the noise level of the noise source for EMC immunity tests. Therefore, there is no option to reduce the noise level for EMC immunity tests.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_12

Fig. 12.1 Noise coupling

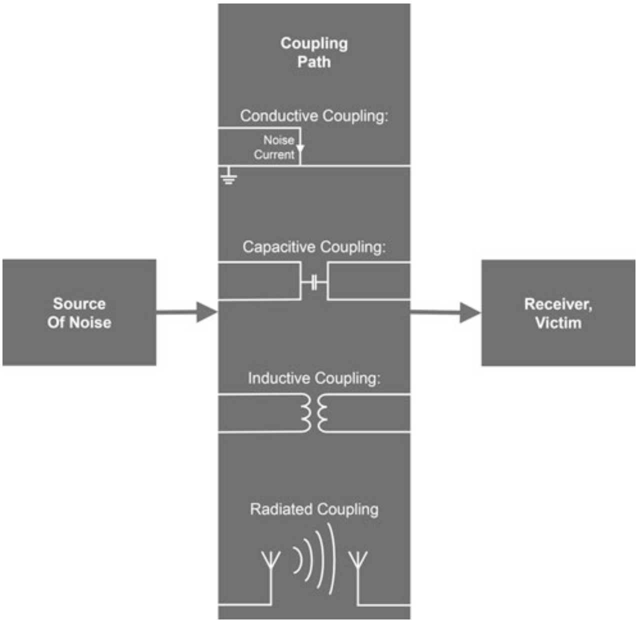

Fig. 12.2 EMI noise coupling paths

2. **Make changes to the coupling path.** The coupling path is where an EMC engineer usually has to focus on if a product fails an EMC test. With well-considered changes to the coupling path, emissions can be reduced and immunity increased (e.g., improving ground connections, adding filters and shields to cables and PCBs).
3. **Increase the victim's immunity level with software or firmware.** Possible software features which can increase a product's EMC immunity are digital filters (e.g., median), spike removers (remove spikes in sensor signals), and sanity checks.

Let us have a closer look. There are different kinds of coupling paths, some of them are conductive (galvanic), and some are nonconductive (radiated). Figure 12.2 shows the four different types of coupling:

1. **Conductive coupling** (common impedance coupling)
2. **Capacitive coupling** (electric field coupling)
3. **Inductive coupling** (magnetic field coupling)

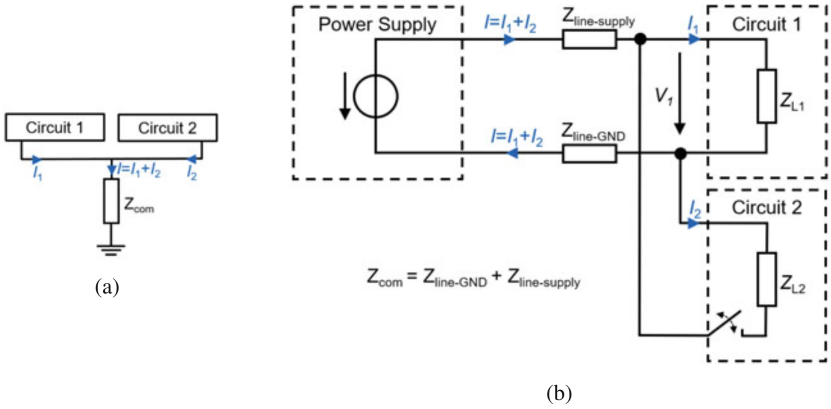

Fig. 12.3 Conductive coupling examples. (**a**) Shared path to ground. (**b**) Shared power supply

4. **Radiated coupling** (electromagnetic field coupling)

The four coupling types are discussed in Sects. 12.1.1–12.1.4.

12.1.1 Conductive Coupling

Conductive coupling can occur when two or more circuits share a common path or common conductor and, therefore, a common impedance $\underline{Z}_{com}$ [Ω]. This shared path leads typically to ground or another reference plane (see Fig. 12.3a). Why could a common conductor path lead to EMC problems or a failed EMC test case? Here is an example: if one of the two circuits in Fig. 12.3a experiences an ESD, burst, or surge pulse, a high current may flow for a short time through a common conductor segment and introduces a noise voltage to the other circuit, which may lead to a malfunction. This is the reason why it is crucial having low-impedance earth/ground planes. A low-impedance ground plane helps to minimize the conductive coupling. In addition, a common path often leads to bad signal integrity as well.

Suppose that circuit 1 in Fig. 12.3a drives some power electronics and circuit 2 is a sensitive measurement circuit. The high current of the power electronics (noise current) introduces a noise voltage $\underline{V}_n$ [V] in the said common impedance $\underline{Z}_{com}$ [Ω] and will lead to interference on the measurement circuit 2. Another example is shown in Fig. 12.3b, where two circuits share a power supply. In the case that circuit 2 is switched off and $\underline{I}_2$ [A] suddenly drops to 0 A, voltage $\underline{V}_1$ [V] will increase by $\Delta\underline{V}_1 = \underline{I}_2\left(\underline{Z}_{line-supply} + \underline{Z}_{line-GND}\right)$.

Generally speaking, the noise voltage $\underline{V}_n$ [V] introduced to the victim circuit is proportional to the common impedance $\underline{Z}_{com}$ [Ω]:

$$\underline{V}_n = \underline{I}_n \cdot \underline{Z}_{com} \tag{12.1}$$

where:

$\underline{I}_n$ = current of the noise source in [A]
$\underline{Z}_{com}$ = common impedance of noise source and victim in [Ω]

Equation 12.1 states that the noise voltage $\underline{V}_n$ [V] is independent of the frequency f [Hz] of the noise signal. However, it must be mentioned that the common impedance $\underline{Z}_{com}(f)$ [Ω] is a function of frequency, and it is to be expected that $|\underline{Z}_{com}(f)|$ [Ω] will increase with increasing frequency (skin effect, inductance L [H] of $\underline{Z}_{com}$). Note: whether the amplitude of the noise voltage $|\underline{V}_n|$ [V] is acceptable or not must be decided under consideration of the useful signal's amplitude (signal-to-noise ratio; see Sect. 6.6).

Some possible measures against conductive coupling are given in the list below. However, be aware that not every option is feasible in any case. The best option has to be decided for every case individually, as there is always a trade-off between project timelines, budget, and hardware costs.

- **Separate current loop.** The best and most effective measure against conductive coupling is separating the noise current loop from the victim's current loop. In case of a complete separation, $\underline{Z}_{com} = 0$.
- **Reduce coupling impedance.** Try to make the coupling impedance as low as possible, e.g., by reducing the inductance (using solid copper planes on the PCB) or reducing the resistance (adding additional copper wires for the ground connection).
- **Filtering.** Adding a filter to the noise source helps prevent undesired noise current, whereas adding a filter to the victim will increase the victim's immunity to noise signals. These filters are typically low-pass filters with capacitors or ferrites.

12.1.2 Capacitive Coupling

Capacitive coupling can occur if there is a coupling capacitance C [F] between two circuits (see Fig. 12.4). The field of concern for the capacitive coupling is the electric E-field. Thus, capacitive coupling is a *near-field coupling* (near-field; see Sect. 8.3), which means that the noise source and the victim, which receives the noise, are closely located to each other.

It is assumed that the capacitive coupling discussed in this section is a weak coupling. That means that the noise coupling is a one-way effect from the noise source to the victim circuit (negligible back reaction from victim to source).

Why could capacitive coupling lead to problems during EMC testing? For example, several wires are together in the same cable, which means that each wire is capacitively coupled to the other wires inside that cable (the capacitance is the

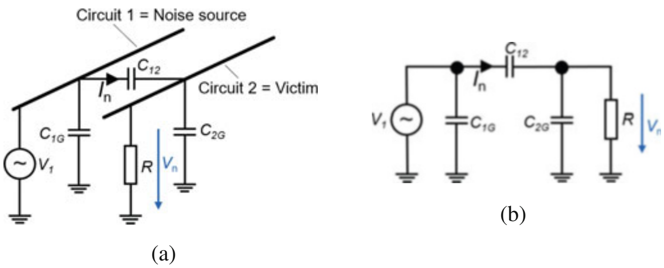

Fig. 12.4 Capacitive coupling between two circuits. (**a**) Physical representation. (**b**) Equivalent circuit

bigger, the longer the cable, and the closer the wires to each other). One of the wires drives the reset signal of a controller system. During EMC testing, an ESD pulse happens to a connector pin that is connected to a wire in that cable. Suppose there are no measures against ESD at the connector pin. In that case, this pulse may discharge through one of the cable's wires and couple capacitively into the other wires in the same cable and potentially resets the controller (if the controller reset signal is not appropriately filtered).

Capacitive coupling happens due to a noise source $v_1(t)$ [V] with high dv_1/dt [V/s] where a noise current $i_n(t)$ [A] is coupled via a stray capacitance C_{12} [F] to the victim. The faster the voltage change dv_1/dt [V/s] and the larger the stray capacitance C_{12} [F], the higher the noise current $i_n(t)$ [A] through C_{12} [F] [6, 7]:

$$i_n(t) = C_{12}\frac{dv_1}{dt} \tag{12.2}$$

The noise voltage $\underline{V}_n$ [V] introduced to the victim circuit of Fig. 12.4 due to a capacitive coupling can be calculated like this [6]:

$$\underline{V}_n = \frac{j\omega\left(\frac{C_{12}}{C_{12}+C_{2G}}\right)}{j\omega + \frac{1}{R(C_{12}+C_{2G})}}\underline{V}_1 \tag{12.3}$$

where:

$\underline{V}_n$ = noise voltage introduced to the victim (circuit 2) in [V]
$\underline{V}_1$ = voltage of the noise source (circuit 1) in [V]
C_{1G} = total capacitance of the noise source circuit to ground in [F]
C_{2G} = total capacitance of the victim circuit to ground in [F]
C_{12} = total couple/stray capacitance between circuit 1 and circuit 2 in [F]
R = load resistor of the victim circuit in [Ω]
$\omega = 2\pi f$ = angular frequency of the sinusoidal noise signal in [rad/sec]

In most cases, R [Ω] is of much lower impedance than the parallel impedances of C_{12} and C_{2G} [6]:

$$R \ll \frac{1}{\frac{1}{1/(j\omega C_{12})} + \frac{1}{1/(j\omega C_{2G})}} = \frac{1}{j\omega\,(C_{12} + C_{2G})} \tag{12.4}$$

If the condition of Eq. 12.4 is given, we can simplify the noise voltage $\underline{V}_n$ [V] calculation to [6]:

$$\underline{V}_n = j\omega R C_{12} \underline{V}_1 \text{ , in case of relatively low } R \tag{12.5}$$

$$v_n(t) = i_n(t) \cdot R = C_{12} \frac{dv_1}{dt} \cdot R \text{ , in case of relatively low } R \tag{12.6}$$

where:

$\underline{V}_n$ = noise voltage introduced to the victim (circuit 2) in [V]
$v_n(t)$ = noise voltage introduced to the victim (circuit 2) in [V]
$\underline{I}_n$ = noise current through coupling capacitance in [A]
$i_n(t)$ = noise current through coupling capacitance in [A]
$\underline{V}_1$ = voltage of the noise source (circuit 1) in [V]
C_{12} = total couple/stray capacitance between circuit 1 and circuit 2 in [F]
R = load resistor of the victim circuit in [Ω]
$\omega = 2\pi f$ = angular frequency of the sinusoidal noise signal in [rad/sec]

In case the resistance R [Ω] in circuit 2 to ground is large compared to the parallel impedances of C_{12} and C_{2G}, such that [6]:

$$R \gg \frac{1}{\frac{1}{1/(j\omega C_{12})} + \frac{1}{1/(j\omega C_{2G})}} = \frac{1}{j\omega\,(C_{12} + C_{2G})} \tag{12.7}$$

then Eq. 12.4 reduces to a capacitive voltage divider [6]:

$$\underline{V}_n = \left(\frac{C_{12}}{C_{12} + C_{2G}} \right) \underline{V}_1 \text{ , in case of relatively large } R \tag{12.8}$$

where:

$\underline{V}_n$ = noise voltage introduced to the victim (circuit 2) in [V]
$\underline{V}_1$ = voltage of the noise source (circuit 1) in [V]
C_{12} = total couple/stray capacitance between circuit 1 and circuit 2 in [F]
C_{2G} = total capacitance of the victim circuit to ground in [F]

This means for a very large R [Ω], the conductive noise coupling is independent of the frequency ω [rad/sec] and the noise voltage is of larger amplitude than when R [Ω] is small. The frequency response in Fig. 12.5 confirms that.

Given all this information, we can now define some possible measures against a capacitive coupling (assuming that the voltage level $\underline{V}_1$ [V] of the noise generator and the victim circuit load R [Ω] cannot be changed):

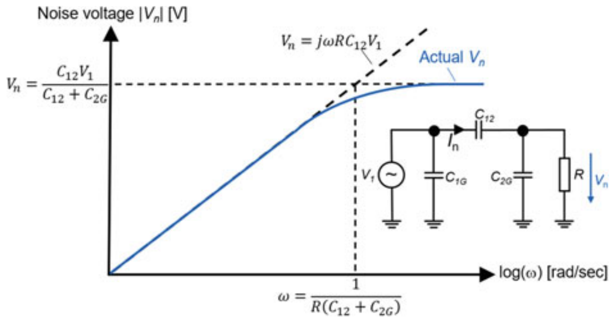

Fig. 12.5 Frequency response of capacitively coupled noise voltage $|\underline{V}_n|$

- **Reduce transient of noise source.** Because the noise current through the stray capacitance $i_n(t) = C_{12}dv_1/dt$ is directly proportional to the voltage transient dv_1/dt [V/s] of the noise source, a reduction of the voltage transient in the noise source helps minimize the capacitive coupling. Note: this measure does not help in the case of very high impedance victim circuits, because in this case, the coupling does not depend on the frequency ω [rad/sec] (see Eq. 12.8).
- **Reduce coupling capacitance.** No coupling capacitance means no capacitive coupling. In practice, the coupling capacitance cannot be removed completely, but it can be minimized with several measures:
  - **Shielding.** Adding a shield around the noise source or the victim reduces the coupling capacitance to a minimum (given that the shield does cover the complete noise source and/or victim). Shielding against capacitive coupling is already achieved by connecting only one end of the shield to ground [6].
  - **Spatial separation.** If possible, the source of noise and the victim circuit could be spatially separated from one another.
  - **Change conductor orientation.** Let us say the noise source circuit and the victim circuit have conductors parallel to each other. If feasible, a change of the orientation to an angle of 90° between the wires would reduce the coupling capacitance.
- **Filtering.** With a filter added to the victim circuit, the coupled noise could be reduced. For high impedance victim circuits, adding a filter capacitor parallel to R [Ω] would increase C_{2G} [F] and therefore help to reduce $|\underline{V}_n|$ [V] (see Eq. 12.8).

12.1.3 Inductive Coupling

Inductive coupling can occur if there is a mutual inductance M [H] between two or more circuits. The field of concern for the inductive coupling is the magnetic

H-field. Thus, inductive coupling is a *near-field coupling* (near-field; see Sect. 8.3), which means that the noise source and the victim, which receives the noise, are closely located to each other.

It is assumed that the inductive coupling discussed in this section is a weak coupling. That means that the noise coupling is a one-way effect from the noise source to the victim circuit (negligible back reaction from victim to source).

How could inductive coupling lead to problems during EMC testing? For example, inductive coupling could lead to issues during EMC testing when high currents flow through cables and PCB traces, e.g., during surge testing where currents up to several [kA] could occur. Large currents often induce voltages in neighboring circuits, which inadvertently disrupts them. These induced voltages are due to inductive coupling.

Inductive coupling happens due to a noise current $i_1(t)$ [A] that changes over time di_1/dt [A/s] and induces a noise voltage $v_n(t)$ [V] in a nearby circuit via a mutual inductance M_{12} [H]. The faster the current change di_1/dt [A/s] and the larger the mutual inductance M_{12} [H], the higher the noise voltage $v_n(t)$ [V] induced into the victim circuit [6]:

$$v_n(t) = M_{12}\frac{di_1}{dt} \tag{12.9}$$

The noise voltage $\underline{V}_n$ [V] induced to the victim circuit shown in Fig. 12.6c due to an inductive coupling can be calculated like this [6]:

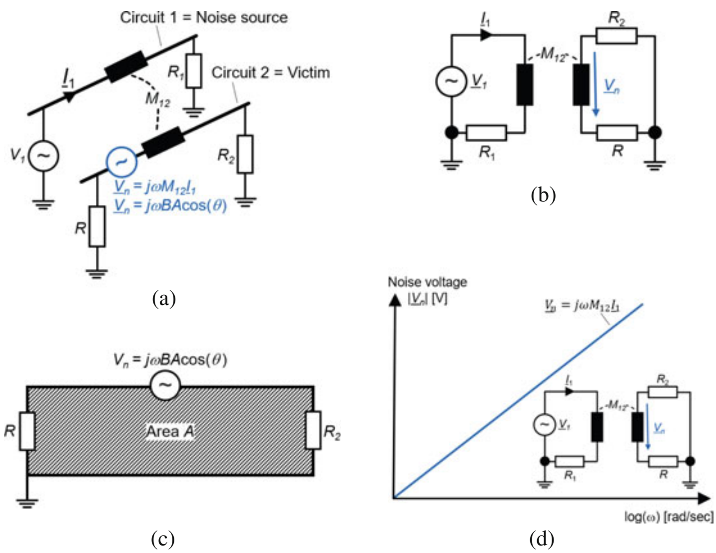

Fig. 12.6 Inductive coupling between two closely spaced circuits. (**a**) Physical representation. (**b**) Equivalent circuit (**c**) Induced noise depends on the area enclosed by the distributed victim circuit. Magnetic field of magnetic flux density B [T] cuts the area A [m^2] at an angle θ [rad]. (**d**) Frequency response of inductive coupled noise voltage $|\underline{V}_n|$ [V]

$$\underline{V}_n = j\omega|B||A|cos\,(\theta) = j\omega\Phi_{12} = j\omega M_{12}|\underline{I}_1| \tag{12.10}$$

where:

$\underline{V}_n$ = induced noise voltage to the victim (circuit 2) in [V]
$|B|$ = magnetic flux density caused by the noise source current (circuit 1) in [T]
$|A|$ = enclosed loop area of the victim circuit (circuit 2) in [m^2]
θ = angle between the magnetic flux vector $\overrightarrow{B}$ and the area vector $\overrightarrow{A}$ in [rad]
Φ_{12} = magnetic flux coupled into the victim circuit area loop (circuit 2) in [Wb]
$M_{12} = \Phi_{12}/|\underline{I}_1|$ = mutual inductance between circuit 1 and circuit 2 in [H]
$\underline{I}_1$ = noise source current (circuit 1) in [A]
$\omega = 2\pi f$ = angular frequency of the sinusoidal noise signal in [rad/sec]

Equations 12.9 and 12.10 show the parameters that influence the induced noise voltage $\underline{V}_n$ [V] to the victim. Suppose that the amplitude of the noise current cannot be changed, we still have the following parameters left, which are all directly proportional to the amplitude of the noise voltage: frequency ω [rad/sec], the magnetic flux density B [T], the area A [m^2], and the angle θ [rad].

Given all this information, we can now define some possible measures against an inductive coupling (assuming that the amplitude of the noise current $\underline{I}_1$ [A] cannot be changed):

- **Reduce current loop area.** The most effective and best way to overcome inductive coupling is to reduce the current loop area A [m^2] of the victim circuit. Therefore, always take care that the current loop of sensitive circuits is kept as small as possible.
- **Spatial separation.** The physical separation of the noise source and the victim circuit leads to a reduced magnetic flux density B [T] (when it reaches the victim current loop).
- **Shielding.** A cable shield for eliminating inductive coupling must be grounded at both ends (a nonmagnetic shield grounded only on one end does not affect a magnetic near-field coupling) [6]. Low-frequency magnetic fields are best shielded with magnetic conductive shielding material ($\mu_r' \ll 1$).
- **Change circuit orientation.** The magnetic flux density is a vector field $\overrightarrow{B}$ [T], which means it has a defined direction. Therefore, a possible option to reduce the noise due to inductive coupling is to change the orientation of the victim current loop area that the term cos (θ) in Eq. 12.10 gets ideally to zero.

12.1.4 Electromagnetic Coupling

Electromagnetic coupling is a *far-field coupling* (far-field; see Sect. 8.3). This means that the noise source and the victim are located far from each other compared to the wavelength λ [m]. The field of concern for radiated coupling is the electromagnetic field (EM-field; see Sect. 8), where the H-field and the E-field travel perpendicular to each other, and their amplitudes both fall off with the factor $1/d$, where d [m] is the distance to the radiating source.

The electromagnetic coupling also plays an important role during EMC tests, e.g., during the radiated immunity test according to IEC 61000-4-3, where the EUT is placed in the far-field of the radiating E-field test antenna (80 MHz to 6 GHz, typically $d = 3$ m), and, moreover, during radiated emission testing according to CISPR 32, where the EUT is the radiator and the receiving antenna is an E-field antenna placed in the far-field (30 MHz to 6 GHz, typically $d = 3$ m or $d = 10$ m).

In the case of an electrical apparatus or installation, any structure may be a good antenna for electromagnetic radiation, e.g., wires, cables, circuit loops or structures on PCBAs, or even larger structures like metal chassis. The best option to develop a robust product against EMI, which produces very low unintended radiation, is always achieved by considering EMC right from the beginning of the project. However, once a product fails EMC testing, there are several possible measures to reduce electromagnetic coupled emissions and increase immunity against electromagnetic fields. Note: because most antennas are reciprocal, a measure against unintended radiation also helps to increase immunity and vice versa.

Here are some typical options on how to reduce unwanted electromagnetic coupling:

- **Filtering.** Suppose a cable is identified as a transmitter or receiver of electromagnetic radiation. In that case, a clamp ferrite bead may reduce the unintended radiation and/or increase the immunity (the clamp ferrite bead attenuates common-mode currents through the cable). Also, filter elements like capacitors placed on the PCB for every IO-line of a cable can help to filter high-frequency signals in either direction. For differential signal interfaces, common-mode chokes are also a good filter option.
- **Shielding.** Put a shield around the circuit, cable, or unit that unintentionally radiates or acts as a receiver of electromagnetic radiation.
- **Grounding.** If a metal chassis or a subsystem within the EUT is the source/sink of the electromagnetic radiation, the ground connection should be checked and improved (low-inductance ground connection). Poor grounding of PCBAs, cable shields, or metal structures often leads to unwanted electromagnetic coupling.

12.2 Differential-Mode vs. Common-Mode

In EMC, a distinction is made between *common-mode* and *differential-mode* noise currents. Both noise currents could lead to unintended radiation, as shown in Sect. 9.9. Sections 12.3 and 12.4 show how the coupling paths presented above can cause differential-mode and common-mode noise. In practice, things are more complicated than stated in this section because coupling mechanisms often happen in cascades and interfere with each other. Nonetheless, you get an idea of where to focus when designing an electrical system or electronic circuit.

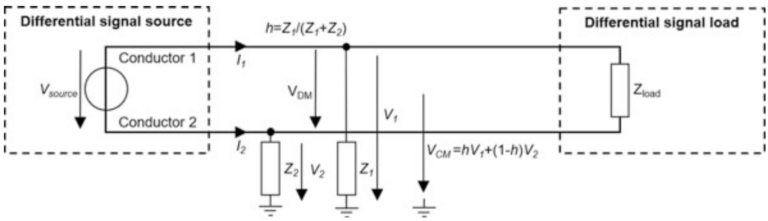

Fig. 12.7 Differential-mode vs. common-mode

Before we jump into the topic of common-mode and differential-mode noise, let us define the terms differential-mode and common-mode in Sects. 12.2.1 (voltages) and 12.2.2 (currents).

12.2.1 Differential-Mode vs. Common-Mode Voltage

Let us have a look at Fig. 12.7, and let us define the differential-mode and common-mode voltages:

- **Differential-mode voltage.** The differential-mode voltage $\underline{V}_{DM}$ [V] is defined as the difference of the voltage potentials to ground of conductor 1 and conductor 2 in Fig. 12.7: $\underline{V}_{DM} = \underline{V}_1 - \underline{V}_2$.
- **Common-mode voltage.** In the case of a balanced transmission line (both conductors have equal impedances to ground), the common-mode voltage $\underline{V}_{CM}$ [V] is 0.5 of the sum of the voltages of conductor 1 and conductor 2 to ground: $\underline{V}_{CM} = (\underline{V}_1 + \underline{V}_2)/2$. However, in the case of an unbalanced transmission line, things are getting a little more complicated, and therefore, we have to define the imbalance factor $\underline{h}$.

For a transmission line with two conductors—like shown in Fig. 12.7,—the *imbalance factor* $\underline{h}$ is defined as [4]:

$$\underline{h} = \frac{\underline{Z}_1}{\underline{Z}_1 + \underline{Z}_2} \tag{12.11}$$

where:

$\underline{h}$ = imbalance factor or current divisor factor of a transmission line
$\underline{Z}_1$ = characteristic impedance of conductor 1 with respect to ground in [Ω]
$\underline{Z}_2$ = characteristic impedance of conductor 2 with respect to ground in [Ω]

The imbalance factor $\underline{h}$ is used for calculating the common-mode voltage $\underline{V}_{CM}$ [V] of any transmission line for a given differential-mode voltage $\underline{V}_{DM}$ [V] [4]:

$$\underline{V}_{DM} = \underline{V}_1 - \underline{V}_2 \tag{12.12}$$

$$\underline{V}_{CM} = \underline{h} \cdot \underline{V}_1 + (1 - \underline{h}) \cdot \underline{V}_2 \tag{12.13}$$

where:

$\underline{V}_{DM}$ = differential-mode voltage between conductor 1 and 2 in [V]
$\underline{V}_{CM}$ = common-mode voltage to ground in [V]
$\underline{V}_1$ = voltage of conductor 1 with respect to ground in [V]
$\underline{V}_2$ = voltage of conductor 2 with respect to ground in [V]
$\underline{h}$ = imbalance factor or current divisor factor of a transmission line

12.2.2 Differential-Mode vs. Common-Mode Current

The difference between *differential-mode* **Differential-Mode Current** and *common-mode* current:

- **Differential-mode current.** A differential-mode current flows in *different* directions through a cable (see Fig. 12.8).
- **Common-mode current.** A common-mode current flows in the same direction—a *common* direction—through a cable (see Fig. 12.13).

Let us have a look at Fig. 12.7. The *imbalance factor* $\underline{h}$ defined in Eq. 12.11 can be used for calculating the differential-mode and common-mode current of any transmission line [4]:

$$\underline{I}_{DM} = (1 - \underline{h}) \cdot \underline{I}_1 - \underline{h} \cdot \underline{I}_2 \tag{12.14}$$

$$\underline{I}_{CM} = \underline{I}_1 + \underline{I}_2 \tag{12.15}$$

where:

$\underline{I}_{DM}$ = differential-mode current in [A]
$\underline{I}_{CM}$ = common-mode current through conductor 1 and 2 in [A]
$\underline{I}_1$ = current through conductor 1 in [A]
$\underline{I}_2$ = current through conductor 2 in [A]
$\underline{h}$ = imbalance factor or current divisor factor of a transmission line

12.3 Differential-Mode Noise Sources

Differential-mode noise sources cannot be found on any bill of material (BOM) or any schematic. These noise sources are unwanted and unintended. Section 9.9.1

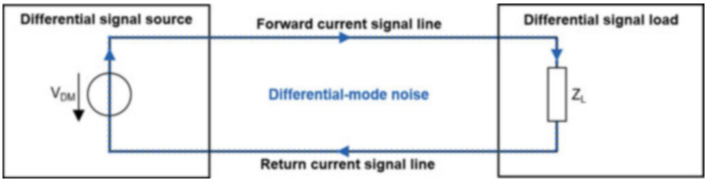

Fig. 12.8 Differential-mode noise current [8]

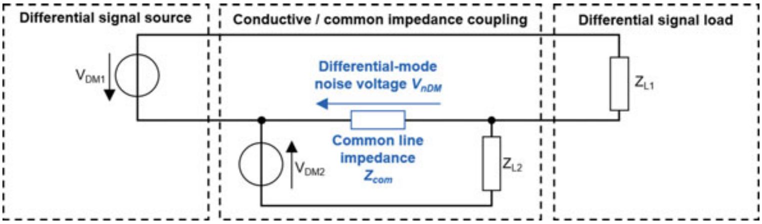

Fig. 12.9 How conductive coupling causes differential-mode noise (simplified)

shows how small differential-mode current loops could lead to unwanted radiated emissions.

A not conclusive list of possible differential-mode interference noise sources is given here and then shortly explained in the following sections:

- **Conductive coupling** to differential-mode noise: Sect. 12.3.1.
- **Capacitive coupling** to differential-mode noise: Sect. 12.3.2.
- **Inductive coupling** to differential-mode noise: Sect. 12.3.3.
- **Common-to-differential-mode conversion**: Sect. 12.3.4.

12.3.1 Conductive Coupling to Differential-Mode Noise

Two or more circuits share a common current path (e.g., a common return current path), and the voltage drop of one of these circuits introduces a differential-mode noise voltage $\underline{V}_{nDM}$ [V] in another circuit(s) and vice versa (see Fig. 12.9). The calculation of the differential-mode noise voltage $\underline{V}_{nDM}$ [V] and more details about conductive coupling—also named *common impedance coupling*—are given in Sect. 12.1.1.

12.3.2 Capacitive Coupling to Differential-Mode Noise

Assuming a noise current $\underline{I}_{nDM}$ [A] is coupled capacitively to one of the two lines of a differential signal. This noise current causes differential-mode noise voltage

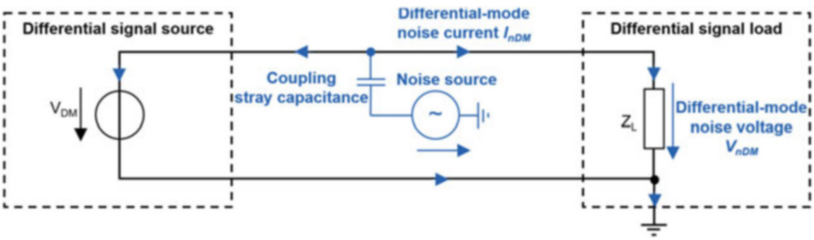

Fig. 12.10 How capacitive coupling to a single conductor causes differential-mode noise (simplified)

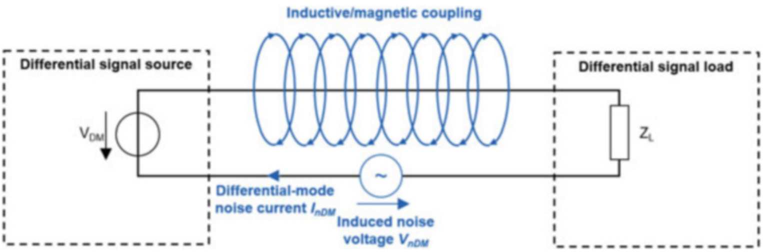

Fig. 12.11 How inductive coupling causes differential-mode noise voltage (simplified)

$\underline{V}_{nDM}$ [V] and could therefore lead to problematic interference on that circuit or unintended RF emissions (see Fig. 12.10). The calculation of the differential-mode noise current $\underline{I}_{nCM}$ [A] and more details about the *capacitive coupling* are given in Sect. 12.1.1.

12.3.3 Inductive Coupling to Differential-Mode Noise

Let's imagine a magnetic field which induces a noise voltage $\underline{V}_n$ [V] to a differential signal current loop which leads to a differential-mode noise current I_n [A] in that circuit (see Fig. 12.11). This induced noise voltage and the noise current could lead to, e.g., problematic interference on that circuit or unintended RF emissions. The calculation of the induced differential-mode noise voltage $\underline{V}_n$ [V] and more details about inductive coupling are given in Sect. 12.1.3.

12.3.4 Common-to-Differential-Mode Conversion

A common-mode noise current $\underline{I}_{nCM}$ [A] can be converted to a differential-mode noise voltage $\underline{V}_{nDM}$ [V] in the case of unbalanced lines. Unbalanced lines mean that

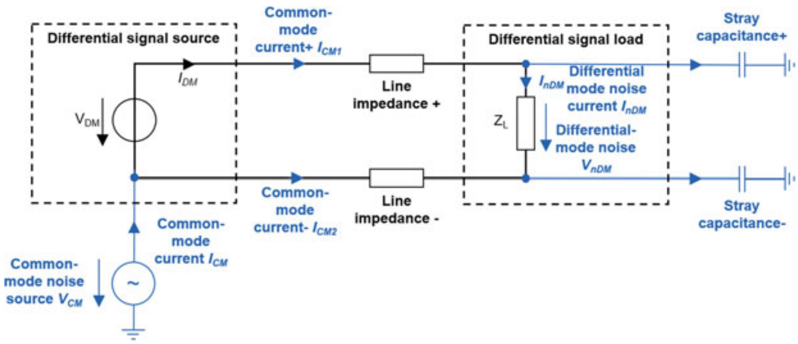

Fig. 12.12 How mode conversion (due to unbalanced lines) causes differential-mode noise voltage (simplified)

the lines are not 100 % balanced;—in other words, the forward and return current lines of a differential signal have [3]:

1. Different impedances along their length
2. Different impedances to ground.

In both cases, a common-mode current leads to a differential-mode voltage (see Fig. 12.12). This type of conversion is called *common-to-differential-mode conversion*.

Electronics designers should especially ensure that signal lines at the input of high-gain instrument amplifiers are well balanced. Otherwise, even a tiny common-mode noise current $\underline{I}_{nCM}$ [A] could cause a significant noise voltage $\underline{V}_{nDM}$ [V] at the output of the high-gain amplifier.

12.4 Common-Mode Noise Sources

Common-mode noise sources cannot be found on any bill of material (BOM) or any schematic. These noise sources are unwanted and unintended. Common-mode currents are the number 1 source of unintentional radiated emissions (Fig. 12.13). Sections 9.9.2, 9.9.3, and 9.9.4 show how common-mode currents in cables could lead to unintended radiated emissions.

A not conclusive list of possible common-mode interference noise sources is given here and then shortly explained in the following sections:

- **Capacitive coupling** to common-mode noise: Sect. 12.4.1.
- **Electromagnetic coupling** to common-mode noise: Sect. 12.1.4.
- **Reference point noise** to common-mode noise: Sect. 12.4.3
- **Differential-to-common-mode conversion**: Sect. 12.4.4.

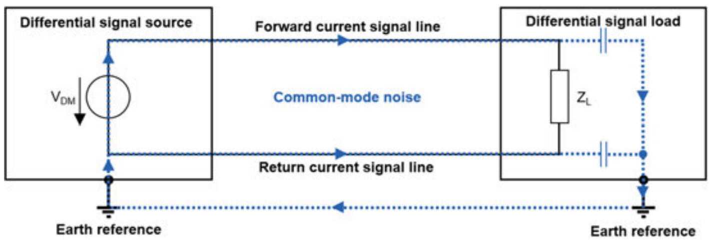

Fig. 12.13 Common-mode noise current [8]

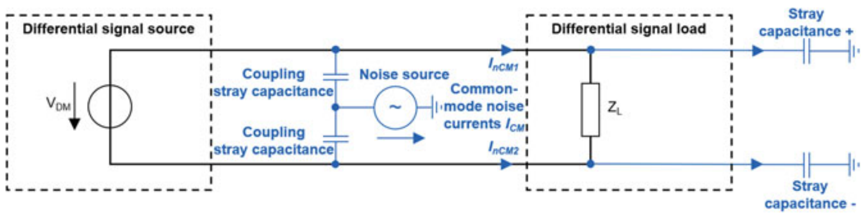

Fig. 12.14 Concurrent capacitive coupling on a pair of wire causes common-mode noise currents, which could lead to high radiated emissions (simplified)

12.4.1 Capacitive Coupling to Common-Mode Noise

Let us imagine that noise currents are coupled capacitively (and equally) onto two differential signal lines, which form a current loop. As a consequence, the coupled common-mode current $\underline{I}_{nCM}$ [A] could lead to radiated or conducted emission issues.

Figure 12.14 shows a simplified setup of how electric near-field coupling could cause common-mode noise. In addition, Sect. 12.1.2 presents the calculation of the coupled noise current and which actions can be taken to reduce it. Finally, how a differential signal leads to common-mode noise with electric field coupling is presented in the following paper [1], where the coupling is called voltage-driven mechanism.

12.4.2 Electromagnetic Coupling to Common-Mode Noise

A cable or a part of an electronic circuit acts as an antenna and receives electromagnetic radiation, e.g., from radio stations or smartphones, etc (Fig. 12.15). The receiving structure is in the far-field of the transmitter. With this coupling, an unintended noise current is brought to an electrical circuit and could lead to interference. Details about the electromagnetic coupling and how to reduce it can be found in Sect. 12.1.4.

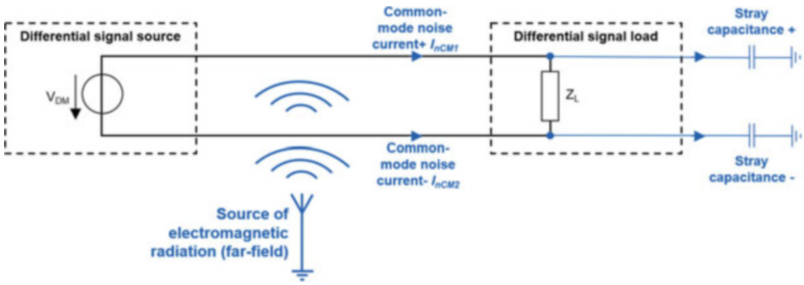

Fig. 12.15 Electromagnetic coupling that leads to a common-mode current (simplified)

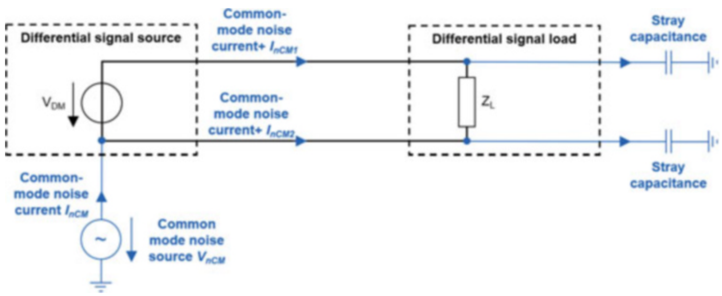

Fig. 12.16 Noisy reference point that leads to a common-mode current (simplified)

12.4.3 Reference Point Noise to Common-Mode Noise

A noisy reference point (ground) or a voltage potential difference of spatially divided circuits and systems could cause a common-mode noise current $\underline{I}_{nCM}$ [A] and, therefore, high radiated emissions. Figure 12.16 shows in a simplified form a setup with a noisy reference point. The noise source voltage $\underline{V}_{nCM}$ [V] itself could be caused by a current-driven differential-to-common-mode conversion [1].

12.4.4 Differential-to-Common-Mode Conversion

Any *change* in electrical balance to ground of a transmission line converts some power of the differential-mode signal into a common-mode signal [4]. This *differential-to-common-mode conversion* happens if—and only if—there is a change in balance of the transmission line [2]. Because this circumstance often causes high radiated emissions and failed EMC tests, the differential-to-common-mode conversion will be analyzed in the following by applying the *imbalance difference modeling* method (IDM [2, 5]).

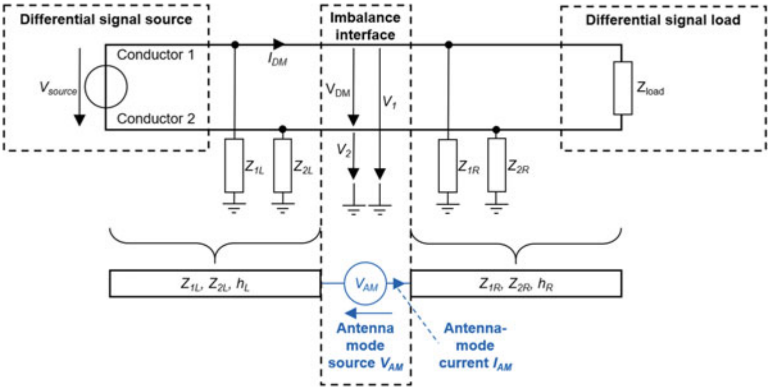

Fig. 12.17 Differential-to-common-mode conversion due to imbalance of a transmission line (change of characteristic impedances to ground)

Consider a transmission line that experiences a sudden change in the impedance of its lines to ground. This is typically the case when the transmission line leads from a PCB to a cable: in the connector, the individual conductors of the transmission line experience a sudden change in the characteristic impedance to ground (change in balance or imbalance). Let us look at Fig. 12.17, where a differential-mode signal experiences a sudden change in impedance to ground (at the imbalance interface). This sudden change leads to a *common-mode voltage* $\underline{V}_{AM}$ [V] and *common-mode current* $\underline{I}_{AM}$ [A], which we call antenna-mode voltage and antenna-mode current because they could lead to unintended radiation. We defined the *imbalance factor* $\underline{h}$ in Eq. 12.11. On the left side of the imbalance interface, conductors 1 and 2 have an imbalance factor of $\underline{h}_L = \underline{Z}_{1L}/(\underline{Z}_{1L} + \underline{Z}_{2L})$, and on the right side of the imbalance interface, the conductors have the imbalance factor of $\underline{h}_R = \underline{Z}_{1R}/(\underline{Z}_{1R} + \underline{Z}_{2R})$. The common-mode voltage $\underline{V}_{AM}$ [V] and current $\underline{I}_{AM}$ [A] that are developed due to the imbalance change are given as [4, 9]:

$$\underline{V}_{AM} = \underline{V}_{DM} \cdot \Delta\underline{h} \tag{12.16}$$

$$\underline{I}_{AM} = \frac{\underline{V}_{AM}}{\underline{Z}_{AM}} \tag{12.17}$$

where:

$\underline{V}_{DM}$ = differential-mode voltage at the imbalance interface in [V]
$\underline{V}_{AM}$ = common-mode voltage developed at the imbalance interface in [V]
$\Delta\underline{h} = \underline{h}_L - \underline{h}_R$ = change of imbalance factor $\underline{h}$ at the imbalance interface
$\underline{I}_{AM}$ = common-mode current through conductor 1 and 2 in [A]
$\underline{Z}_{AM}$ = input impedance of antenna that is formed by conductors 1 and 2 in [Ω]

Imbalance difference modeling (IDM) is a powerful method for quickly calculating the worst-case common-mode voltage driving a cable (as an antenna). As a worst-case, it can be assumed an imbalance factor change of $\Delta h = 0.5$ (perfectly balanced to completely unbalanced and vice versa), which means that the developed common-mode voltage is half the differential-mode amplitude:

$$|\underline{V}_{AM}| = |\underline{V}_{DM}| \cdot 0.5 \tag{12.18}$$

In EMC, the theoretical maximum field strength E_{max} [V/m] for a given setup at distance d [m] in the far-field is of primary interest. Therefore, here is a way for a worst-case estimation of the radiated emissions of a digital signal due to transmission line imbalance (e.g., change from an unbalanced microstrip line to a balanced ribbon cable):

1. **Differential-mode voltage $\mathbf{V_{DM}}$**. Use the Fourier analysis to determine the RMS amplitude V_{DM} [V] of the sinusoidal harmonics of differential-mode signal. See Sect. M.2 for an estimation of harmonics of an ideal square wave signal.
2. **Common-mode voltage $\mathbf{V_{AM}}$**. Calculate the assumed worst-case common-mode RMS voltage V_{AM} [V] developed at the transmission line imbalance change for the harmonics of interest [4]:

$$V_{AM} = 0.5 \cdot V_{DM}$$

3. **Worst-case radiated emissions $\mathbf{E_{AM,max}}$**. Use Eq. 9.48 on page 130 to calculate the estimated worst-case radiated emissions E_{max} [V/m] for the given distance d [m] (far-field) and the antenna-mode voltage V_{AM} [V] (current, voltage, and field-strength are RMS values):

$$E_{AM,max} \approx \begin{cases} \frac{60 \cdot I_{AM,max}}{d} \cdot \frac{2}{\sin\left(\sqrt{2}\right)} & \text{when } l_{cable} \le \frac{\lambda}{2} \\ \frac{60 \cdot I_{AM,max}}{d} \cdot \frac{2}{\sin\left(\sqrt{\frac{\lambda}{l_{cable}}}\right)} & \text{when } l_{cable} > \frac{\lambda}{2} \end{cases} \tag{12.19}$$

where:

$$\underline{I}_{AM,max} = \frac{\underline{V}_{AM}}{\frac{36.5\,\Omega}{k_{board} \cdot k_{cable}}} \tag{12.20}$$

$$k_{board} = \begin{cases} \sin\left(\frac{2\pi \cdot l_{board}}{\lambda}\right) & \text{when } l_{board} \le \frac{\lambda}{4} \\ 1.0 & \text{otherwise} \end{cases} \tag{12.21}$$

$$k_{cable} = \begin{cases} \sin\left(\frac{2\pi \cdot l_{cable}}{\lambda}\right) & \text{when } l_{cable} \le \frac{\lambda}{4} \\ 1.0 & \text{otherwise} \end{cases} \tag{12.22}$$

$$l_{board} = \frac{1 + \frac{2L}{W}}{\frac{2L}{W}} \cdot \sqrt{L^2 + W^2} \tag{12.23}$$

V_{AM} = common-mode RMS voltage driving the monopole antenna (cable) in [V]
$I_{AM,max}$ = highest common-mode RMS current that exists on the cable in [A]
$R_{min} = 36.5\,\Omega$ = radiation resistance of a resonant $\lambda/4$-monopole antenna in [Ω]
l_{board} = effective board length in [m]
l_{cable} = cable length in [m]
L = length of the PCBA (board) in [m]
W = width of the PCBA (board) in [m]
k_{board} = impact of board size l_{board} on R_{min} in case $l_{board} \leq \lambda/4$
k_{cable} = impact of cable length l_{cable} on R_{min} in case $l_{cable} \leq \lambda/4$
d = distance to the radiating structure (PCBA, cable) in [m]
λ = wavelength of the sinusoidal signal $\underline{V}_{AM}$ in [m]

12.5 Summary

Table 12.1 presents a summary of EMI noise coupling paths.

Table 12.1 EMI noise coupling path summary

Coupling path description	Category	Typical noise source	Predominant physical quantity	How to typically reduce coupling path effectiveness?
Common Impedance Coupling	Galvanic	High current circuts (source) and sensitive circuits (victim)	**Common impedance $\underline{Z}$**	**Separate common current paths** of noise source and victim. **Reduce coupling impedance**. Add filter to circuit of source or victim (e.g. capacitor, ferrite)
Capacitive Coupling	Near-field	Fast signal transients, high voltage	**Electric field, E-field**	**Reduce transient of noise source** signal. **Shielding** of noise source or victim. **Spatial separation** of source and victim. Add filter to victims circuit (e.g. capacitor, ferrite)
Inductive Coupling	Near-field	High currents, inductors, transformers	**Magnetic field, H-field**	**Reduce current loop area of victim**. **Shielding** of noise source or victim. **Spatial separation** of source and victim. Add filter to victims circuit (capacitor, ferrite)
Radiated Coupling	Far-field	Radiators (wireless devices, antennas), PCB traces and cables	**Electromagnetic field**	**Shielding** of noise souce or victim. Add **filter** to victim/radiator circuit (ferrite bead, common-mode choke, capacitor). Improve **grounding**

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Chapter 13
Shielding

Action expresses priorities.

—Mahatma Gandhi

In the field of EMC, shields are used to:

1. **Reduce electromagnetic emissions** from a product.
2. **Increase immunity** against electric, magnetic, and/or electromagnetic radiation.

The shielding theory presented in this book is based on the accepted shielding theory for electromagnetic waves, initially proposed by Schelkunoff ((1943) Electromagnetic waves. D. van Nostrand Company Inc, New York, pp 303–312) in 1943. The formulas in this chapter are approximations for shields with high electrical conductivity. Before we jump into the theory of shielding, here are two practical pieces of advice:

1. **Cables and wires.** Every single signal which enters and/or leaves a shielded enclosure must be filtered or shielded. In case the cable is shielded, contact the cable shield 360° with the shielded enclosure.
2. **Slots and apertures.** Slots and apertures reduce the shielding effectiveness SE or even lead to higher emissions than without the shield in case of resonances inside a shielding enclosure Hubing ([1] EMC Question of the Week: 2017–2020. LearnEMC, LLC, Stoughton). If the linear dimension l [m] of a slot or aperture is larger than $\lambda/2$, the shield is assumed to be useless Ott ([3] Electromagnetic compatibility engineering. Wiley, New York).

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
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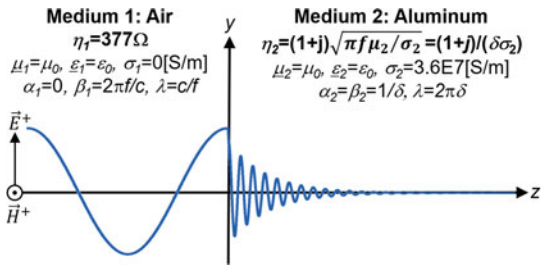


Fig. 13.1 Attenuation of a sinusoidal electromagnetic wave due to absorption loss. The electromagnetic wave comes from the left and impinges on a thick metal shield in the far-field (symbolic drawing; reflection is not shown here)

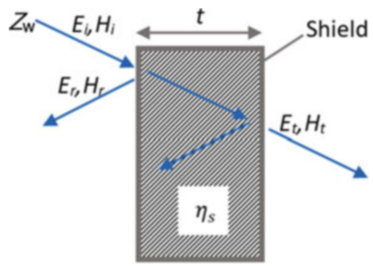


Fig. 13.2 Shield with intrinsic impedance $\underline{\eta}_s$ [Ω] and thickness t [m]. Reflection happens at the left boundary (primarily for the E-field) and the right boundary (primarily for the H-field) [4]

13.1 Shielding Theory

Shielding of electromagnetic waves is usually achieved by:

- **Reflection.** Let us suppose an electromagnetic wave that impinges on a shield with an intrinsic impedance lower than the wave impedance: $|\underline{\eta}_s| < |\underline{Z}_w|$. In that case, the E-field is partially reflected at the shield's outer surface, and the H-field is partially reflected at the inner surface (see Fig. 13.2) [5].
- **Absorption.** When an electromagnetic wave propagates through a lossy media (like a good conductor with $\alpha > 0$), the amplitudes of the electric E-field and the magnetic H-field decrease exponentially with $e^{-\alpha t}$, where α [1/m] is the attenuation coefficient and t [m] the shield thickness (see Fig. 13.1).

13.2 Shielding Effectiveness

Let us have a look at Fig. 13.2 and let us define a few terms. The *shielding effectiveness* (SE) describes how well a shield attenuates an incident wave (electrical field strength E_i [V/m], magnetic field strength H_i [A/m]) when propagating through a shield. After passing through a shield, the remaining wave has a field strength of E_t [V/m] and H_t [A/m]. The reflected wave has field strength E_r [V/m] and H_r [A/m]. Shielding effectiveness is now defined as [4]:

$$\mathrm{SE}_{E,dB} = 20\log_{10}\left|\frac{E_i}{E_t}\right| \tag{13.1}$$

$$\mathrm{SE}_{H,dB} = 20\log_{10}\left|\frac{H_i}{H_t}\right| \tag{13.2}$$

where:

$|E_i|$ = electric field strength of the incident wave to the shield barrier in [V/m]
$|E_t|$ = electric field strength of the wave when it leaves the shield barrier in [V/m]
$|H_i|$ = magnetic field strength of the incident wave to the shield barrier in [A/m]
$|H_t|$ = magnetic field strength of the wave when leaving the shield in [A/m]
$\mathrm{SE}_{E,dB}$ = shield effectiveness for the E-field in [dB]
$\mathrm{SE}_{H,dB}$ = shield effectiveness for the H-field in [dB]

The shielding effectiveness of the E-field and the H-field compared:

- $\mathbf{SE_{E,dB} = SE_{H,dB}}$. In the case of a uniform plane wave and identical media on both sides of the shield barrier, the shielding effectiveness for the E-field and the H-field are equivalent [4].
- $\mathbf{SE_{E,dB} \neq SE_{H,dB}}$. In the case of near-fields and/or different media on both sides of the shield barrier, the shielding effectiveness of the E-field and the H-field are not equivalent [4] because of the different reflection losses R_{dB} [dB].

The shielding effectiveness is the sum of *reflection loss* R_{dB} [dB], *absorption loss* A_{dB} [dB], and the *multiple-reflection loss correction* M_{dB} [dB] (Fig. 13.9).

$$\mathrm{SE}_{dB} = \underbrace{R_{dB}}_{\substack{Reflection\\ loss}} + \underbrace{A_{dB}}_{\substack{Absorption\\ loss}} + \underbrace{M_{dB}}_{\substack{Multiple-\\ reflection\\ loss\\ correction}} \tag{13.3}$$

Assuming a uniform plane wave arriving perpendicular to a solid shield, which does not have any slots or apertures, we can write [4]:

$$R_{dB} = 20\log_{10}\left|\frac{\left(\underline{Z}_w + \underline{\eta}_s\right)^2}{4\underline{Z}_w\underline{\eta}_s}\right| \tag{13.4}$$

$$A_{dB} = 20\log_{10}\left|e^{\alpha t}\right| \tag{13.5}$$

$$M_{dB} = 20\log_{10}\left|1 - \left(\frac{\underline{Z}_w - \underline{\eta}_s}{\underline{Z}_w + \underline{\eta}_s}\right)^2 e^{-2\alpha t}\right| \tag{13.6}$$

where:

R_{dB} = reflection loss of both boundaries of the shield (attenuation due to reflection of the incident wave at the boundaries of the shield, E-fields are primarily reflected at the first boundary—when entering the shield, whereas H-fields are primarily reflected at the second boundary—when leaving the shield) in [dB]

A_{dB} = absorption loss (attenuation due to power converted to heat as the wave propagates through the shield, absorption loss is more critical for H-fields than for E-fields because the H-field reflection loss primarily happens at the second boundary when leaving the shield) in [dB]

M_{dB} = multiple-reflection loss correction (applicable for very thin shields $t < \delta$, where absorption loss is low and therefore the energy of the H-field at the second boundary is still significant) in [dB]

$\underline{Z}_w = \eta_0 \approx 377\,\Omega$ wave impedance of the incident wave in [Ω]

$\underline{\eta}_s$ = intrinsic impedance of the shield material in [Ω]

t = thickness of the shield in [m]

α = attenuation coefficient of the shield material in [1/m]

Looking at the equations above, the following can be said:

- **Reflection loss R.**
  - R dominates at lower frequencies because the absorption loss A is low.
  - R in the near-field differs with changing radiation source impedance.
  - R in the near-field is different from R in the far-field.
- **Absorption loss A.**
  - A dominates at higher frequencies, where the skin depth is small.
  - A is identical for near-field and far-field radiation.
- **Multiple-reflection loss correction M.**
  - M can be neglected for good conducting ($|\eta_s| \ll |Z_w|$) and thick shields ($t \gg \delta$), where $A_{dB} < 15\,\mathrm{dB}$.
  - M must be considered for thin shields ($t \ll \delta$), where multiple reflections inside the shield reduce the shielding effectiveness.

13.3 Far-Field Shielding

Figure 13.3 presents the absorption, reflection, and multiple-reflection loss correction of a solid shield in the *far-field*. It can be seen how the reflection loss dominates for low frequencies and the absorption loss for high frequencies.

13.3.1 Reflection Loss R for Far-Field Shielding

The reflection loss R_{dB} [dB] of a solid shield with good conductivity ($|\underline{\eta}_s| \ll |\eta_0|$) in the far-field can be approximated as:

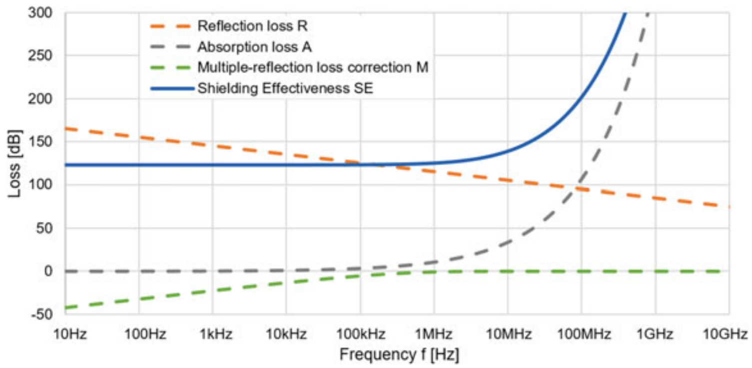


Fig. 13.3 Shielding effectiveness of a solid aluminum shield with $t = 0.1$ mm in the far-field

$$R \approx \left| \frac{\eta_0}{4\underline{\eta}_s} \right| = \sqrt{\frac{\sigma}{4\pi f \mu_r' \varepsilon_0}} \tag{13.7}$$

$$R_{dB} \approx 20 \log_{10} \left| \frac{\eta_0}{4\underline{\eta}_s} \right| = 20 \log_{10} \left[\sqrt{\frac{\sigma}{4\pi f \mu_r' \varepsilon_0}} \right] \tag{13.8}$$

where:

R = reflection loss for a plane wave and a solid shield with good conductivity
$\eta_0 = \mu_0/\varepsilon_0 \approx 377\,\Omega$ = intrinsic impedance of free-space in [Ω]
$|\underline{\eta}_s| = |(1+j)\sqrt{\pi f \mu_0 \mu_r'/\sigma}| = \sqrt{2\pi f \mu_0 \mu_r'/\sigma}$ = shield impedance in [Ω]
σ = specific conductance of the shield material in [S/m]
f = frequency of the sinusoidal plane wave in [Hz]
μ_r' = relative magnetic permeability of the shield material
$\varepsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum, absolute permittivity

13.3.2 Absorption Loss A for Far-Field Shielding

The absorption loss A_{dB} [dB] of a solid shield barrier of thickness t [m] and of good conductivity ($\alpha = 1/\delta$) in the far-field can be approximated as:

$$A \approx e^{\alpha t} = e^{t/\delta} \tag{13.9}$$

$$A_{dB} \approx 20 \log_{10} e^{t/\delta} = 8.7 \cdot \frac{t}{\delta} = 8.7 \cdot t \cdot \sqrt{\pi f \mu_0 \mu_r' \sigma} \tag{13.10}$$

where:

A = absorption loss for a plane wave and a solid shield with good conductivity
$\alpha = 1/\delta$ = attenuation coefficient of the shield material in [1/m]

t = thickness of the shield in [m]
$\delta = 1/\sqrt{\pi f \mu_0 \mu_r' \sigma}$ = skin depth for a wave with frequency f in [m]
f = frequency of the sinusoidal plane wave in [Hz]
μ_r' = relative magnetic permeability of the shield material
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum, absolute permeability
σ = specific conductance of the shield material in [S/m]

13.3.3 Multiple-Reflection Loss Correction M for Far-Field Shielding

The multiple-reflection loss correction M_{dB} [dB] must be considered for thin shields, where the absorption loss A_{dB} [dB] is below 15 dB. For a plane wave in the far-field and a solid shield barrier of good conductivity, M_{dB} [dB] can be approximately calculated as [3]:

$$M \approx 1 - e^{-2t/\delta} \tag{13.11}$$

$$M_{dB} \approx 20 \log_{10}\left(1 - e^{-2t/\delta}\right) \tag{13.12}$$

where:

M = multiple-reflection loss correction for a plane wave and a thin solid shield with good conductivity
t = thickness of the shield in [m]
$\delta = 1/\sqrt{\pi f \mu_0 \mu_r' \sigma}$ = skin depth for a wave with frequency f in [m]

13.4 Near-Field Shielding

Figure 13.4 shows that the reflection loss R_{dB} [dB] for electric dipole antennas in the *near-field* is much higher compared to the far-field and compared to magnetic dipoles in the near-field.

Comparing *near-field* vs. *far-field* shielding, we can say:

- **Reflection loss.** R_{dB} [dB] is affected by the wave impedance $\underline{Z}_w$ [Ω] (see Eq. 13.4). Therefore, the reflection loss R_{dB} [dB] for near-fields is different from far-fields and different for low impedance sources (see Sect. 13.4.2) compared to high impedance sources (see Sect. 13.4.1).
- **Absorption loss.** A_{dB} [dB] is unaffected by the wave impedance $\underline{Z}_w$ [Ω] (see Eq. 13.5). Therefore, the absorption loss A_{dB} [dB] is identical for near-field and far-field shielding.

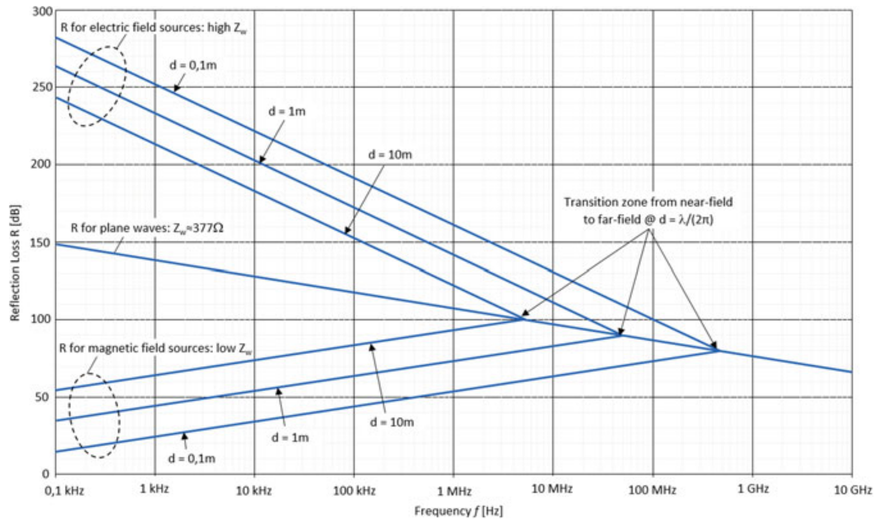


Fig. 13.4 Reflection loss R_{dB} [dB] for various wave impedances $\underline{Z}_w$ [Ω]

- **Multiple-reflection loss correction.** M_{dB} [dB] is affected by the wave impedance $\underline{Z}_w$ [Ω] (see Eq. 13.6). However, the influence of the wave impedance on M_{dB} [dB] is only minor. Therefore, in most cases, it can be assumed that M_{dB} [dB] is not affected significantly by the wave impedance.

13.4.1 Near-Field Shielding of Electric Sources

To calculate the approximate absorption loss A_{dB} [dB] and multiple-reflection loss correction M_{dB} [dB], Eqs. 13.10 and 13.12 can be used. $R_{dB,E}$ [dB] is the reflection loss when shielding a near-field wave from an electric dipole (predominant E-field) and it is approximately:

$$R_e \approx \left| \frac{\underline{Z}_{we}}{4\underline{\eta}_s} \right| \tag{13.13}$$

$$R_{dB,e} \approx 20 \log_{10} \left| \frac{\underline{Z}_{we}}{4\underline{\eta}_s} \right| = 244 + 10 \log_{10} \left(\frac{\sigma}{f^3 \varepsilon_{rw}'^2 \mu_{rs}' d^2} \right) \tag{13.14}$$

where:

$|\underline{Z}_{we}| = 1/(2\pi f \varepsilon_{rw}' \varepsilon_0 d)$ = near-field wave impedance of E-field antenna in [Ω]
$|\underline{\eta}_s| = |(1+j)\sqrt{\pi f \mu_0 \mu_r' / \sigma}| = \sqrt{2\pi f \mu_0 \mu_r' / \sigma}$ = shield impedance in [Ω]
σ = specific conductance of the shield material in [S/m]

f = frequency of the sinusoidal plane wave in [Hz]
μ'_{rs} = relative magnetic permeability of the shield material
ε'_{rw} = relative electric permittivity of the medium through which the near-field wave is propagating (can be assumed to be 1.0 in most cases)
d = distance from the source antenna (electric dipole) to the shield surface in [m]

The near-field to far-field boundary and wave impedances are discussed in Sects. 8.3 and 8.5.

13.4.2 Near-Field Shielding of Magnetic Sources

To calculate the approximate absorption loss A_{dB} and multiple-reflection loss correction M_{dB} [dB], Eqs. 13.10 and 13.12 can be used. $R_{dB,M}$ [dB] is the reflection loss when shielding a near-field wave from a magnetic dipole (predominant H-field) and it is approximately:

$$R_m \approx \left| \frac{\underline{Z}_{wm}}{4\underline{\eta}_s} \right| \tag{13.15}$$

$$R_{dB,m} \approx 20\log_{10}\left| \frac{\underline{Z}_{wm}}{4\underline{\eta}_s} \right| = -63 + 10\log_{10}\left(\frac{\sigma f d^2 \mu'^2_{rw}}{\mu'_{rs}} \right) \tag{13.16}$$

where:

$|\underline{Z}_{wm}| = 2\pi f \mu'_{rw}\mu_0 d$ = near-field wave impedance of H-field antenna in [Ω]
$|\underline{\eta}_s| = |(1+j)\sqrt{\pi f \mu_0 \mu'_r/\sigma}| = \sqrt{2\pi f \mu_0 \mu'_r/\sigma}$ = shield impedance in [Ω]
σ = specific conductance of the shield material in [S/m]
f = frequency of the sinusoidal plane wave in [Hz]
μ'_{rs} = relative magnetic permeability of the shield material
μ'_{rw} = relative magnetic permeability of the medium through which the near-field wave is propagating (can be assumed to be 1.0 in most cases)
d = distance from the source antenna (electric dipole) to the shield surface in [m]

Note: Eq. 13.16 is an approximation and assumes $|\underline{Z}_{wm}| \gg |\underline{\eta}_s|$. In case of a negative reflection loss $R_{dB,m}$, set $R_{dB,m} = 0$ and $M_{dB_m} = 0$. In case $R_{dB,m}$ is positive and near zero, Eq. 13.16 is in slight error [3].

The near-field to far-field boundary and wave impedances are discussed in Sects. 8.3 and 8.5.

13.4.3 Low-Frequency Magnetic Field Shielding

Low-frequency ($f < 100\,\text{kHz}$) magnetic field waves are the most difficult waves to shield because the absorption loss is low due to the low-frequency f [Hz] and the reflection loss is low due to the low wave impedance $\underline{Z}_w$ [Ω]. Figure 13.5 illustrates this problem for a solid 1 mm aluminum shield in the far-field and the near-field.

Therefore, two additional methods for *shielding low-frequency magnetic fields* are presented here:

- **Shields with $\boldsymbol{\mu'_r \gg 1.0}$.** Use a shielding material with high magnetic permeability $\mu'_r \gg 1.0$ (e.g., mu-metal; see Sect. G.1). The drawing in Fig. 13.6 should illustrate how a low-reluctance shield will divert the magnetic flux in an environment with $\mu'_r = 1$. Note: magnetic permeability decreases with increasing frequency f [Hz] and with increasing magnetic field strength H [A/m] or magnetic flux Φ [Wb], respectively.
- **Shorted turn method.** Use a loop conductor which is placed in the magnetic field H [A/m]. The induced current $\underline{I}_{ind}$ [A] in the loop conductor will generate a counter magnetic field that leads to a reduced magnetic field in the vicinity of the loop (see Fig. 13.7).

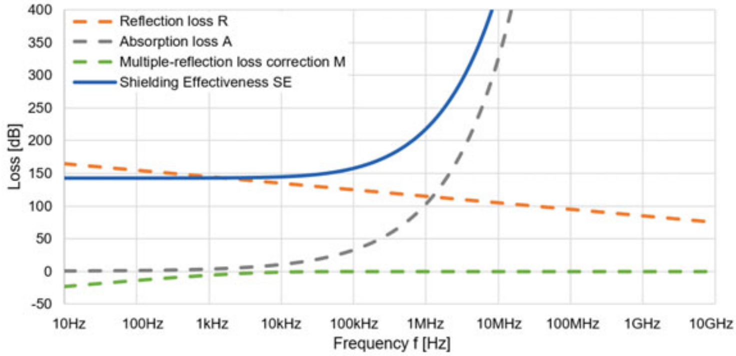
13.5 Slots and Apertures

Slots and apertures can cause considerable leakage if their largest dimension is $l > (\lambda/10)$. In addition, they are efficient radiators (yes, radiators!) when their maximum linear dimension l [m] is equal to $\lambda/2$ [2]. Thus, if a slot or aperture has a linear dimension of $l \geq \lambda/2$, the shielding effectiveness SE_{dB} [dB] can be assumed to be 0 dB.

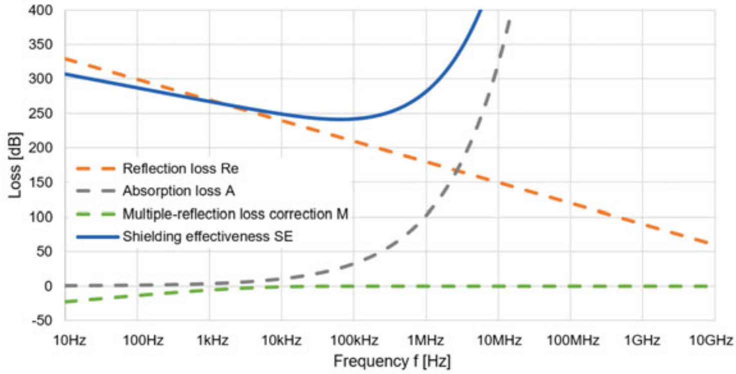
At high frequencies ($f > 1\,\text{MHz}$), slots and apertures are usually of more concern than the attenuation or reflection loss of a shield, because the sum of attenuation and reflection loss of any reasonably thick ($t \gg \delta$) and conductive shield ($|\underline{\eta}_s| \ll |\underline{Z}_w|$) is higher than 100 dB (see Fig. 13.5). In other words, in cases where the intrinsic SE of a shielding material is high, the SE is defined by the slots and apertures of the shield.

13.5.1 Single Aperture

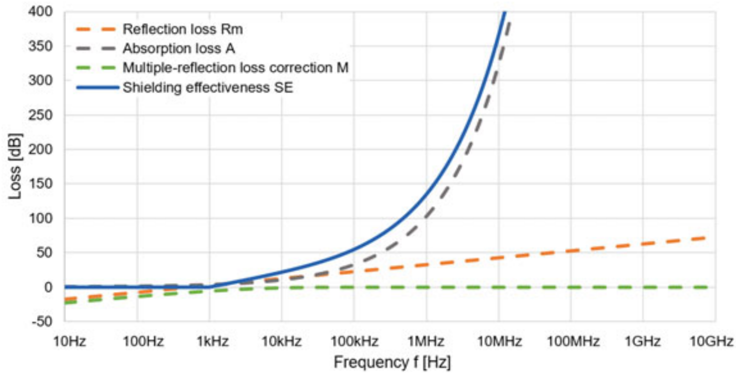
If we assume a shield with high intrinsic SE, so that the SE is defined by the slots and apertures of the shielding enclosure, the SE can be calculated based on the maximum linear dimension l [m] (which must be less than $\lambda/2$). In the case of one single aperture, we can write [3]:



(a)



(b)



(c)

Fig. 13.5 Approximate shielding effectiveness of a solid aluminum shield with thickness $t = 1\,\text{mm}$. (**a**) In the far-field. (**b**) In the near-field of an electric dipole at distance $d = 10\,\text{mm}$. (**c**) In the near-field of a magnetic dipole at distance $d = 10\,\text{mm}$

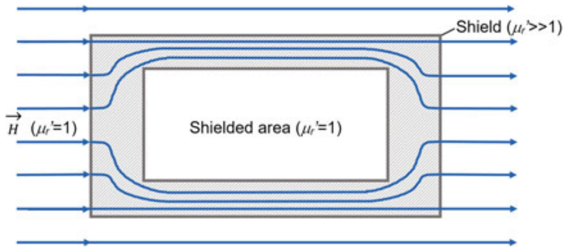


Fig. 13.6 Shielding of low-frequency magnetic field with high-permeability shield material (symbolic)

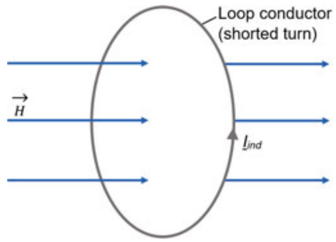


Fig. 13.7 Shielding of low-frequency magnetic field with a shorted turn (conductor loop)

$$\mathrm{SE}_{dB} \approx 20 \log_{10}\left(\frac{\lambda}{2l}\right) \tag{13.17}$$

where:

λ = wavelength of the sinusoidal wave in [m]
l = maximum linear dimension of one single aperture ($l < \lambda/2$) in [m]

13.5.2 Multiple Apertures

If there is not only one aperture but multiple apertures—as shown in Fig. 13.8b—the shielding effectiveness will be reduced even further. The shielding effectiveness of a linear array of equally and closely spaced apertures n of maximum length l [m]—where the total array length l_{array} [m] is less than $\lambda/2$—is approximately [3]:

$$\mathrm{SE}_{dB} \approx 20 \log_{10}\left(\frac{\lambda}{2l}\right) - 20 \log_{10} \sqrt{n} = 20 \log_{10}\left(\frac{\lambda}{2l\sqrt{n}}\right) \tag{13.18}$$

where:

λ = wavelength of the sinusoidal wave in [m]
l = maximum linear dimension of one single aperture with $l < \lambda/2$ in [m]
n = number of closely spaced, identical apertures in an array with $l_{array} < \lambda/2$

In the case of a multidimensional array of m rows (and $m < n$; see Fig. 13.8c), the additional rows (second, third, . . .) will not reduce the SE significantly [3].

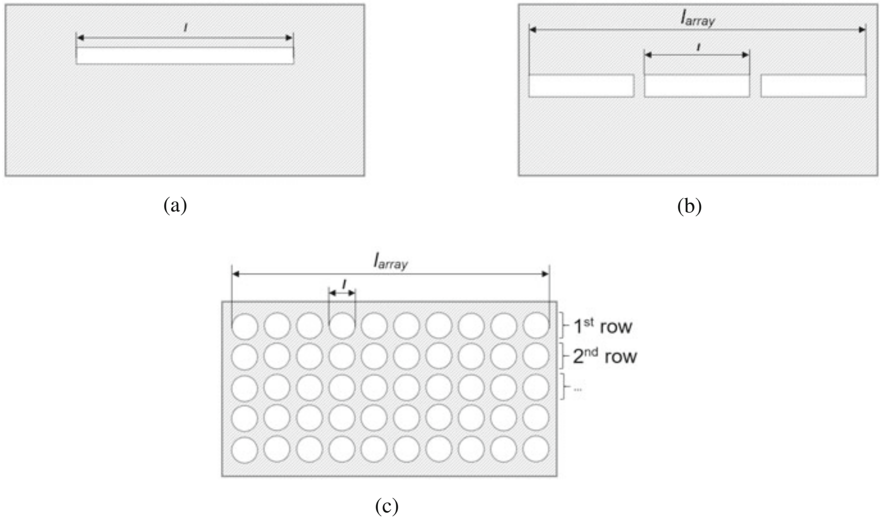
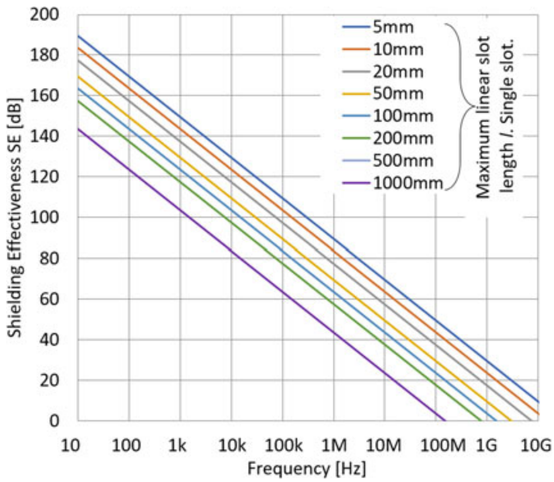


Fig. 13.8 Apertures in shields. (**a**) Single aperture of maximum linear dimension l [m]. (**b**) Linear array of identical and closely spaced apertures. (**c**) Multidimensional array of identical and closely spaced apertures

Fig. 13.9 Shielding effectiveness in the case of a single slot ($l < (\lambda/2)$) and high intrinsic SE of the shield material



Thus, the SE of a multidimensional array of equal-sized apertures is the SE of one single hole, minus the SE reduction of the first row of n apertures (see Eq. 13.18, Fig. 13.10).

Apertures located on different surfaces, which all look in different directions, do not decrease the overall shielding effectiveness because they radiate in different directions [3]. Thus, it is a good idea to distribute apertures around the surface of a shielding enclosure.

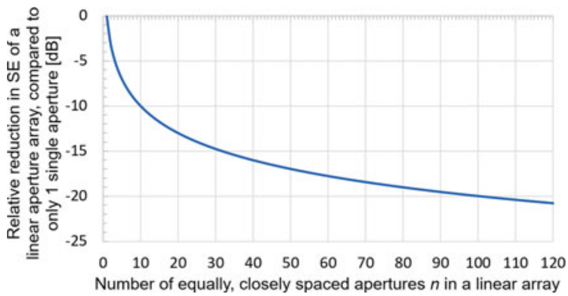


Fig. 13.10 Shielding effectiveness reduction of an array of n identical and closely spaced apertures in comparison with one single aperture

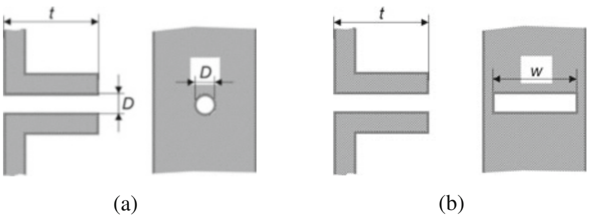


Fig. 13.11 Apertures shaped to form a waveguide. (**a**) Circular aperture with waveguide. Left: cross-sectional view. Right: frontal view. (**b**) Rectangular aperture with waveguide. Left: cross-sectional view. Right: frontal view

13.5.3 Waveguide Below Cutoff

The shielding effectiveness of an aperture can be improved by adding a depth to the aperture. Figure 13.11 shows apertures, which have been extended to *waveguides*. The concept is as follows: the waveguide has a defined cutoff frequency f_c [Hz], and when a wave with frequency $f \ll f_c$ enters the waveguide, it will be attenuated. Above the cutoff frequency f_c [Hz], the waveguide is very efficient at passing all frequencies and thus does not improve the SE of the aperture anymore.

First, we have to determine the cutoff frequency f_c [Hz] of the waveguide aperture, which is given for free-space (air, where $v \approx c$) as [3, 4]:

$$f_{c,cwg} \approx \frac{1.8412 \cdot v}{\pi D} = \frac{1.8412 \cdot c}{\pi D} \tag{13.19}$$

$$f_{c,rwg} \approx \frac{v}{2w} = \frac{c}{2w} \tag{13.20}$$

where:

$f_{c,cwg}$ = waveguide cutoff frequency of a circular waveguide in [Hz]
$f_{c,rwg}$ = waveguide cutoff frequency of a rectangular waveguide in [Hz]
v = velocity of the wave in [m/sec]
$c = 1/(\sqrt{\mu_0 \varepsilon_0}) = 2.998 \cdot 10^8$ m/sec = speed of light
w = largest inner side length (width) of rectangular waveguide in [m]

D = inner diameter of circular waveguide in [m]

Now, we can calculate the shielding effectiveness SE_{dB} of n identical and closely spaced waveguide apertures in case the wave has a frequency $f \ll f_c$ [4] [3]:

$$SE_{dB,cwg} \approx 20 \log_{10}\left(\frac{\lambda}{2l\sqrt{n}}\right) + 32\frac{t}{w} \tag{13.21}$$

$$SE_{dB,rwg} \approx 20 \log_{10}\left(\frac{\lambda}{2l\sqrt{n}}\right) + 27.3\frac{t}{D} \tag{13.22}$$

where:

$SE_{dB,cwg}$ = shielding effectiveness of a circular waveguide aperture in [dB]
$SE_{dB,rwg}$ = shielding effectiveness of a rectangular waveguide aperture in [dB]
λ = wavelength of the sinusoidal wave in [m]
l = maximum linear dimension of one single aperture ($l < \lambda/2$) in [m]
w = largest inner side length (width) of rectangular waveguide in [m]
D = inner diameter of circular waveguide in [m]
t = depth of the waveguide in [m]
n = number of closely spaced, identical apertures in an array with $l_{array} < \lambda/2$ (set $n = 1$ in case of a single aperture)

13.6 Grounding of Shields

First, a disclaimer: this section is about solid shielding enclosures and not about shielded cables. Grounding of cable shields is discussed in Sect. 13.7.4.

A solid *shielding enclosure* does not need to be grounded to be an effective shield for electromagnetic waves from outside of the enclosure to the inside, and vice versa [3]. However, metallic shields are often grounded for safety reasons and to protect the inside from electrostatic charges.

In case cables are leaving or entering the shield, consider these points:

- **Unshielded cables.** The shielding enclosure and the circuit ground must be connected with low impedance close to where the cable leaves and/or enters the enclosure. This way, the common-mode voltage which may drive the cable conductors is reduced to a minimum. In addition, every signal which enters and/or leaves a shielded enclosure should be low-pass filtered.
- **Shielded cables.** The cable shield should be connected with low impedance to the outside of the shielding enclosure. In this case, the cable shield can be considered as the extension of the shielded enclosure.

Consider the galvanic series to prevent corrosion when connecting dissimilar metals for grounding (e.g., aluminum and steel; see Table H.2). This is especially important in the case of safety grounds.

13.7 Cable Shields

13.7.1 Transfer Impedance Z_t

The *transfer impedance* $\underline{Z}_t$ [Ω/m] is a property of *shielded cables*, and it is used to measure the shielding effectiveness of cable shields. The transfer impedance describes the magnetic coupling between the noise current $\underline{I}$ [A] along the shield to the induced voltage $\underline{V}$ [V] to the inner conductor per length l [m]. The smaller the transfer impedance $\underline{Z}_t$ [Ω/m], the more effective the shield.

Figure 13.12 shows the schematic diagram of a transfer impedance measurement setup of a cable shield. $\underline{Z}_t$ [Ω/m] can be calculated as [6]:

$$\underline{Z}_t = \frac{\underline{V}}{l \cdot \underline{I}} \tag{13.23}$$

where:

$\underline{V}$ = measured voltage between the center conductor and the inner surface of the shield in [V]
$\underline{I}$ = test current in the shield in [A]
l = length of the shielded cable under test in [m]

The transfer impedance $\underline{Z}_t(f)$ [Ω/m] is a function of frequency f [Hz]. At low frequencies f [Hz], $\underline{Z}_t \cdot l$ is equal the direct current resistance R_{DC} [Ω] of the shield. At high frequencies ($>$ 1 MHz), $\underline{Z}_t$ [Ω/m] of a solid shield decreases because of the skin effect (current flows at outside layer of the shield), whereas for braided shields, $\underline{Z}_t$ [Ω/m] increases at high frequencies (see Fig. 13.13). Remember: a low transfer impedance $\underline{Z}_t$ [Ω/m] means high shielding effectiveness.

13.7.2 Cable Shielding Against Capacitive Coupling

Figure 13.14 shows a setup of two closely spaced circuits, where circuit 1 is the noise source and circuit 2 is the victim. The noise is capacitively coupled from circuit 1 to circuit 2, where circuit 2 is shielded and the shield is grounded. Whether

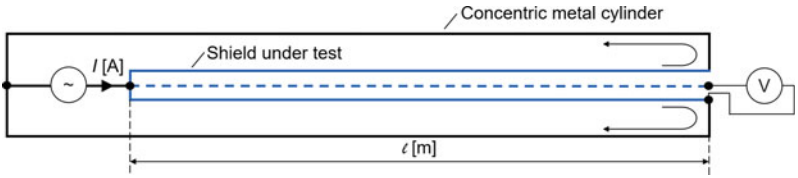


Fig. 13.12 Transfer impedance $\underline{Z}_t$ [Ω/m] measurement circuit

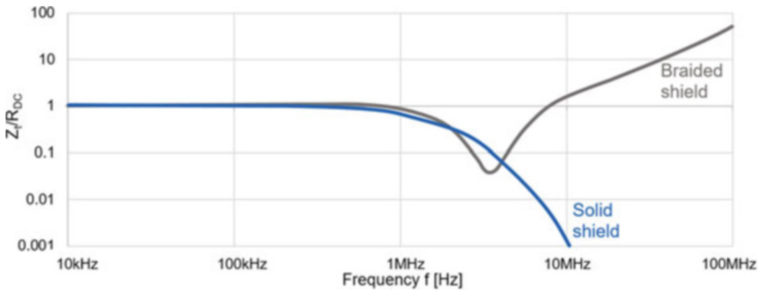
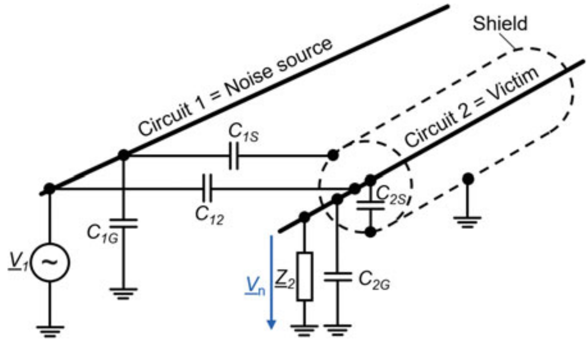


Fig. 13.13 Typical normalized transfer impedance of a solid shield and a braided shield

Fig. 13.14 Simplified circuit representation of cable shielding against capacitive coupling



the shield is grounded at both ends or just at one end does not matter in this case, because the shielding effectiveness against capacitively coupled noise does not depend on a closed-loop shield current; the shielding effectiveness against capacitively coupled noise does mainly depend on the impedance of the shield to ground (the lower the impedance, the better), the value of the stray coupling capacitance C_{12} [F] (the lower, the better), and the victim's impedance $|\underline{Z}_2|$ (the lower, the better).

In the case of a setup like shown in Fig. 13.14, where the shield is grounded with very low impedance, the noise voltage $\underline{V}_n$ [V] can be calculated as:

$$\underline{V}_n = \underline{V}_1 \frac{\frac{1}{\frac{1}{\underline{Z}_2} + j\omega C_{2S} + j\omega C_{2G}}}{\frac{1}{j\omega C_{12}} + \frac{1}{\frac{1}{\underline{Z}_2} + j\omega C_{2S} + j\omega C_{2G}}} = \underline{V}_1 \frac{j\omega C_{12}\underline{Z}_2}{1 + j\omega \underline{Z}_2 (C_{2G} + C_{2S})} \tag{13.24}$$

where:

$\underline{V}_1$ = voltage of the noise source (circuit 1) in [V]
$\underline{Z}_2$ = impedance of the victim circuit to ground in [Ω]
C_{1G} = capacitance of circuit 1 (noise source) to ground (neglected here, because C_{1G} is parallel to the ideal voltage source $\underline{V}_1$) in [F]
C_{2G} = capacitance of circuit 2 (victim) to ground in [F]

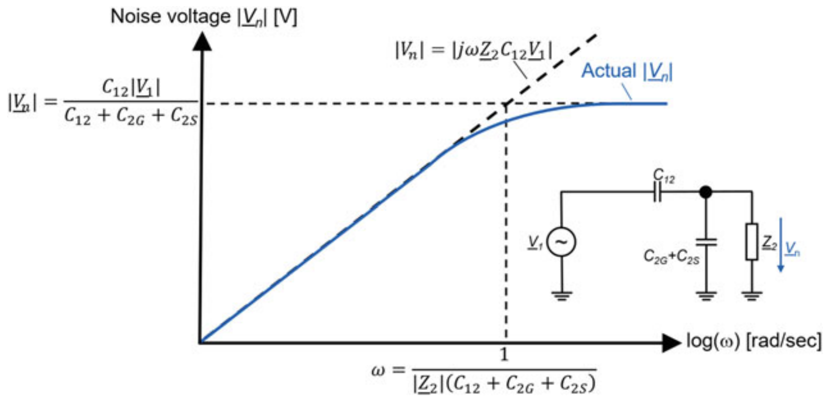


Fig. 13.15 Frequency response of capacitively coupled noise voltage $|\underline{V}_n|$ [V] to a shielded cable (where the shield is a good conductor and grounded)

C_{1S} = capacitance of circuit 1 (noise source) to ground (neglected here, because C_{1S} is parallel to the ideal voltage source $\underline{V}_1$ in case the shield is grounded) in [F]

C_{2S} = capacitance of circuit 2 (victim) to ground (parallel to $\underline{Z}_2$ in case the shield is grounded) in [F]

C_{12} = coupling stray capacitance between circuit 1 and circuit 2 in [F]

Equation 13.24 and Fig. 13.15 lead us to the conclusion that effective shielding against capacitive coupling is all about minimizing the coupling stray capacitance C_{12} [F]. Thus, the following two points are important to remember when it comes to shielding against electric field coupling:

- **Shield grounding.** The cable shield must be grounded with low impedance. Otherwise, the total coupling capacitance would increase from $C_{coupling} = C_{12}$ to $C_{coupling} = C_{12} + 1/(1/C_{1S} + 1/C_{2S})$.
- **Minimize protruding conductors.** In order to minimize the coupling capacitance C_{12} [F], the length of the shielded conductors that extend beyond the shield must be minimized.

13.7.3 Cable Shielding Against Inductive Coupling

Figure 13.16 shows a setup of two closely spaced circuits, where circuit 1 is the noise source and circuit 2 is the victim. The noise is inductively coupled from circuit 1 to circuit 2, where circuit 2 is shielded, and the shield is grounded at both ends. If the shield is made of a nonmagnetic material and the shield would only be connected to ground at one end, there would be no shielding effect, because the

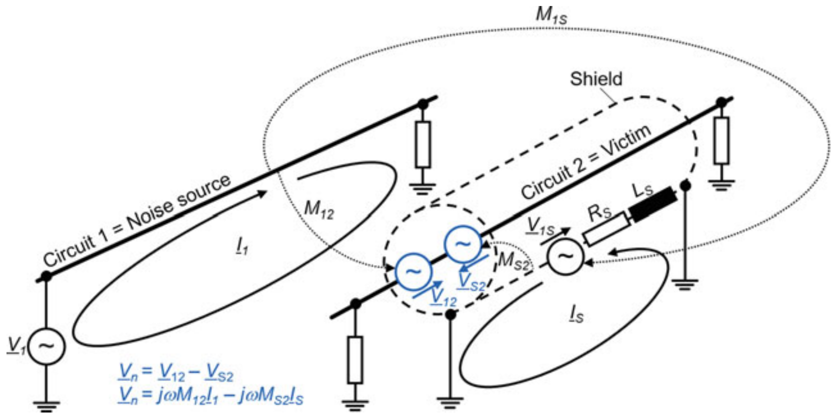


Fig. 13.16 Simplified circuit representation of cable shielding against inductive coupling

shielding effectiveness against inductively coupled noise depends on the induced shield current $\underline{I}_S$ [A] (in the case of a nonmagnetic shield material).

The shielding effect due to the induced shield current is explained like this:

1. **Noise voltage induction.** Current $\underline{I}_1$ [A] of the noise source induces via mutual inductance M_{12} [H] a noise voltage $\underline{V}_{12}$ [V] in the conductor 2.
2. **Shield current induction.** Current $\underline{I}_1$ [A] of the noise source induces via mutual inductance M_{1S} [H] a voltage $\underline{V}_{1S}$ [V] in the shield. This voltage $\underline{V}_{1S}$ [V] leads to the shield current $\underline{I}_S[A]$, which flows along the cable shield and through the ground.
3. **Noise voltage compensation.** The induced shield current $\underline{I}_S$ [A] induces a so-called compensation voltage $\underline{V}_{S2}$ [V] in the conductor 2. Ideally, the compensation voltage $\underline{V}_{S2}$ [V] has the opposite polarity to the induced noise voltage $\underline{V}_{12}$ [V]. Thus, the induced shield current $\underline{I}_S$ [A] leads to a shielding effect.

The induced noise voltage $\underline{V}_{12}$ [V] from circuit 1 (noise source) to circuit 2 (victim) is:

$$\underline{V}_{12} = j\omega M_{12}\underline{I}_1 \tag{13.25}$$

where:

$M_{12} = \Phi_{12}/i_1(t)$ = mutual inductance between circuit 1 and circuit 2 in [H]
$\underline{I}_1$ = current of the noise source (circuit 1) in [A]
$\omega = 2\pi f$ = angular frequency of the sinusoidal noise signal in [rad/sec]

The induced shield voltage $\underline{V}_{1S}$ [V] from circuit 1 (noise source) to the shield—which must be grounded at both ends—is:

$$\underline{V}_{1S} = j\omega M_{1S}\underline{I}_1 \tag{13.26}$$

where:

$M_{1S} = \Phi_{1S}/i_1(t) =$ mutual inductance between circuit 1 and the shield in [H]
$\underline{I}_1 =$ current of the noise source (circuit 1) in [A]
$\omega = 2\pi f =$ angular frequency of the sinusoidal noise signal in [rad/sec]

If we model the shield as simple RL-series-circuit, we can write the shield current $\underline{I}_S$ [A], induced by the noise current $\underline{I}_1$ [A] from circuit 1, as:

$$\underline{I}_S = \frac{\underline{V}_{1S}}{R_S + j\omega L_S} \tag{13.27}$$

where:

$\underline{V}_{1S} =$ induced shield voltage to shield due to $\underline{I}_1$ in circuit 1 (noise source) in [V]
$R_S =$ resistance of the shield current loop in [Ω]
$L_S =$ inductance of the shield current loop in [H]
$\omega = 2\pi f =$ angular frequency of the sinusoidal noise signal in [rad/sec]

The mutual inductance M_{S2} [H] between the shield and any inner conductor is equal to the shield inductance L_S [H] [3]. Therefore, the induced compensation voltage $\underline{V}_{S2}$ [V] to the shielded conductor 2 due to the shield current $\underline{I}_1$ [A] can be written as:

$$\underline{V}_{S2} = j\omega M_{S2}\underline{I}_S = j\omega L_S\underline{I}_S \tag{13.28}$$

where:

$M_{1S} =$ mutual inductance between the shield and any inner conductor in [H]
$L_S =$ inductance of the shield current loop in [H]
$\underline{I}_S =$ induced shield current due to current $\underline{I}_1$ in circuit 1 in [A]
$\omega = 2\pi f =$ angular frequency of the sinusoidal noise signal in [rad/sec]

If the mutual inductance M_{12} [H] of circuit 1 to circuit 2 is equal to the mutual inductance M_{1S} [H] of circuit 1 to the shield around circuit 2 (because the loop area and orientation of the shield and the inner conductors are nearly identical), we can calculate the total induced noise voltage $\underline{V}_n$ [V] in circuit 2 (victim) as:

$$\underline{V}_n = \underline{V}_{12} - \underline{V}_{S2} = j\omega M_{12}\underline{I}_1 - j\omega M_{S2}\underline{I}_S = j\omega\left(M_{12}\underline{I}_1 - L_S\underline{I}_S\right) \tag{13.29}$$

$$= j\omega\left(M_{12}\underline{I}_1 - L_S\frac{\underline{V}_{1S}}{R_S + j\omega L_S}\right) = j\omega\left(M_{12}\underline{I}_1 - \frac{j\omega M_{1S}\underline{I}_1}{R_S/L_S + j\omega}\right) \tag{13.30}$$

$$= j\omega\underline{I}_1\left(M_{12} - \frac{j\omega M_{12}}{R_S/L_S + j\omega}\right) = j\omega M_{12}\underline{I}_1\left(\frac{R_S/L_S}{R_S/L_S + j\omega}\right) \tag{13.31}$$

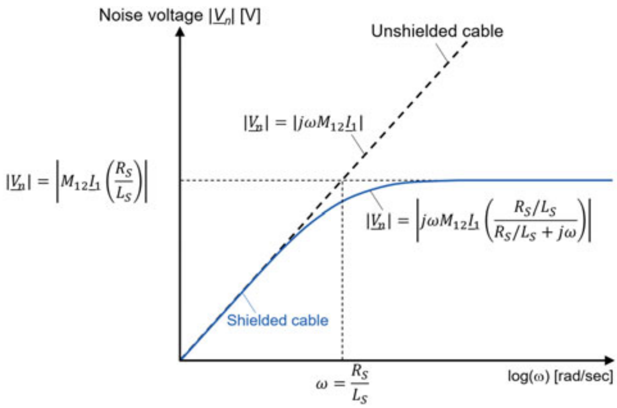


Fig. 13.17 Frequency response of inductively coupled noise voltage $|\underline{V}_n|$ [V] to a shielded cable vs. an unshielded cable (where the shield is a good conductor, nonmagnetic, and grounded at both ends)

where:

$M_{12} = \Phi_{12}/i_1(t)$ = mutual inductance between circuit 1 and circuit 2 in [H]
$\underline{I}_1$ = current of the noise source (circuit 1) in [A]
R_S = resistance of the shield current loop in [Ω]
L_S = inductance of the shield current loop in [H]
$\omega = 2\pi f$ = angular frequency of the sinusoidal noise signal in [rad/sec]

Equation 13.31 and Fig. 13.17 lead us to the conclusion that shielding against magnetic field coupling is all about minimizing the mutual inductance M_{12} [H] and minimizing the shield current loop resistance R_S [Ω] (given the shield is connected at both ends of the cable and is of nonmagnetic material):

- **Minimizing $\mathbf{M_{12}}$**. The mutual inductance M_{12} [H], which leads to the magnetic field coupling, can be minimized most effectively by reducing the current loop area of the victim's circuit (see Sect. 12.1.3). This is why differential signals along a twisted pair (minimal current loop area) are robust against magnetic field coupling.
- **Minimizing $\mathbf{R_S}$**. Minimizing R_S [Ω] means minimizing the shield resistance, the shield termination resistance (connection to ground), and any resistance in the ground loop. Therefore, for maximum shielding effectiveness against magnetic field coupling, all resistances in the shield current loop must be minimized.

Equation 13.31 shows that in the case of nonmagnetic shielding material ($\mu_r' = 1$), there is no shielding effect for low-frequency magnetic fields and that the shielding effectiveness increases with increasing frequency (in case the shielding is grounded at both ends).

13.7.4 Grounding of Cable Shields

To be most effective, cable shields should be connected to the outside of the shielded enclosure if available or to circuit ground otherwise. The connection(s) of the shield to chassis/ground should be of low impedance. The most effective termination is a 360° connection of the shield. This means that the shield is contacted around the entire circumference to ground (e.g., with a conductive cable clamp). Avoid pigtail ground connections because they are of high impedance at high frequencies and noise can be picked up by pigtails.

Question: connecting a cable shield to ground at both ends of the cable or only at one end? Answer: it depends:

- **Shield grounded at one end.** Grounding the shield at only one end prevents low-frequency ($f < 100\,\text{kHz}$) noise current to flow along the shield, which could couple noise to the inner conductors. However, at higher frequencies, a current flow cannot be prevented effectively due to the stray capacitance of the shield to other conductors and ground.

  In addition, grounding at only one end leads to the fact that the shield could act as an antenna for high frequencies where the shield is an effective antenna (e.g., at $l = \lambda/4$). Therefore, shield grounding at only one end is only recommended for low-frequency applications with short cables, where the primary goal is the protection against electric E-fields.
- **Shield grounded at both ends.** Whenever in doubt, connect the cable shield at both ends to chassis/ground. Shield grounding at both ends is recommended for high-frequency ($f > 100\ kHz$), digital circuit applications and when the cable length l [m] is longer than $l > \lambda/20$, where λ [m] is the wavelength of the highest significant frequency of the functional signal inside the cable. In addition, grounding at both ends acts as an additional measure against magnetic H-field coupling, because the magnetic field induces a current to the shield, which induces a voltage in the shielded conductor, which ideally cancels out the induced noise voltage (see Sect. 13.7.3).

  In the case of a solid shield, noise currents with frequency $f > 1\ MHz$ flow primarily at the shield's outer surface. This effect leads to the circumstance that a coaxial cable acts like a triaxial cable (a cable with two shields), where the signal return current flows at the inner surface of the shield and the noise current at the outer surface. This means that there is nearly any common impedance coupling in coaxial cables for frequencies above $f > 1\ MHz$.
- **Hybrid shield grounding.** Figure 13.18 shows the principle of the *hybrid cable shield grounding*. This method of grounding enables a selective current flow through the shield. For high-frequency signals, both sides are connected to ground, which enables the high-frequency current to flow through the shield, and due to the skin effect, the high-frequency current flows at the outer surface of the shield. Thus, the inner conductor is protected against high-frequency

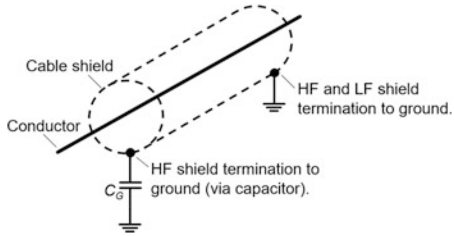


Fig. 13.18 Hybrid grounding of a cable shield

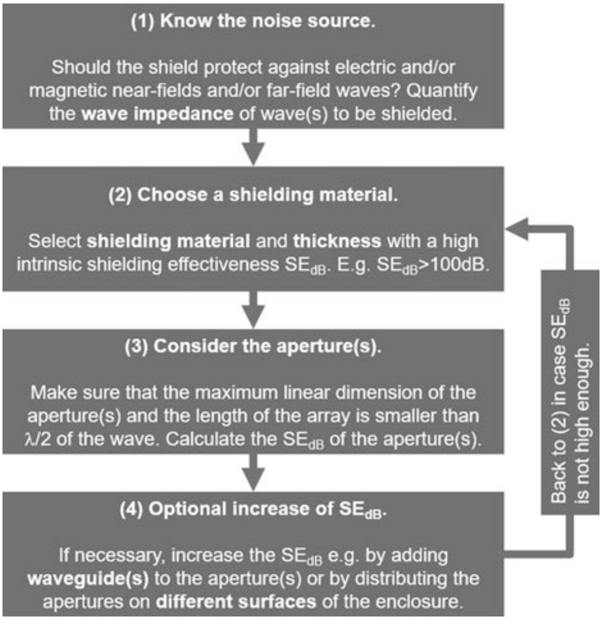


Fig. 13.19 Proposed workflow for designing a shielded enclosure

magnetic fields. On the other hand, low-frequency signals (DC or mains power frequency signals of 50 Hz or 60 Hz) are blocked by the capacitor C_G [F] in Fig. 13.18. Hybrid cable shield grounding is a good option to minimize DC and low-frequency shield currents in case of high voltage differences at both ends of the cable while maintaining a good SE against high-frequency noise.

13.8 Summary

Figure 13.19 presents a proposal for the design workflow of a shielded enclosure. Table 13.1 compares cable shield terminations at one end vs. both ends.

Table 13.1 Termination of cable shields to ground: one end vs. both ends

	Cable shield connected to chassis/ground on one side	Cable shield connected to chassis/ground on both sides
Advantage	- No **low-frequency** shield current - Good SE for low-frequency **electric fields**	- Good SE for **high-frequencies** - Cable shield does not act as an **antenna**
Disadvantage	- Shield could act as **antenna** (e.g. $l=\lambda/4$) - Low SE against **magnetic fields**	- **Low-frequency noise current** in shield may couple noise voltage to the conductors

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Chapter 14
Grounding

Of all the conductors used in interconnecting electronics, the most complex is ironically the one that generally gets the least attention: the ground.

—Henry W. Ott

14.1 Types of Ground

In EMC, a distinction is made between these types of ground:

- **Safety ground.** Safety ground means the ground of an electrically conductive chassis or electric circuit, which protects humans against shock hazards and the electric circuit from ESD pulses.
- **Functional ground.** A ground for other purposes than electrical safety, e.g., for mitigating EMC and EMI issues.
- **Chassis ground.** A chassis ground refers to the connection that establishes an electrical link to a metallic enclosure. For example, cable shields are usually connected to chassis ground.
- **Signal ground.** Signal ground means the return current path of a signal current back to its source.

More ground types and their symbols are mentioned in Sect. 14.2.

14.2 Ground Symbols

As mentioned above in Sect. 14.3, a distinction is made between different ground types. The standard IEC 60417 [1] defines the symbols for safety, functional, and chassis ground. These ground symbols are presented in Fig. 14.1.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_14

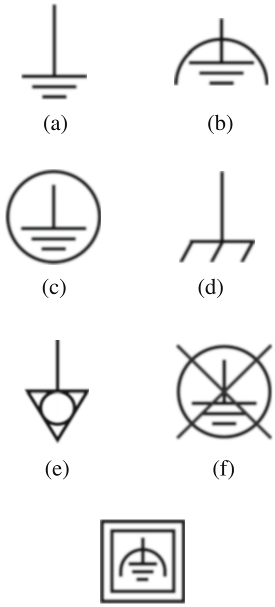


Fig. 14.1 Earth and ground symbols according to IEC 60417. (**a**) No. 5017. Earth, ground. To identify an earth (ground) terminal in cases where neither the symbol 5018 nor 5019 is explicitly required [1]. (**b**) No. 5018. Functional earthing/grounding (US). To identify a functional earthing (grounding) terminal, for example, of a specially designed earthing (grounding) system to avoid causing malfunction of the equipment [1]. (**c**) No. 5019. Protective earth/ground (US). To identify any terminal which is intended for connection to an external conductor for protection against electric shock in case of a fault or the terminal of a protective earth (ground) electrode [1]. (**d**) No. 5020. Frame or chassis. To identify the frame or chassis terminal [1]. (**e**) No. 5021. Equipotentiality. To identify the terminals which, when connected together, bring the various parts of an equipment or of a system to the same potential, not necessarily being the earth (ground) potential, e.g., for local bonding [1]. (**f**) No. 6032. Do not connect to protective earth/ground (US). To indicate that conductive parts shall not be connected to protective earth (ground), e.g., on electrical equipment with conductive parts inside an insulating enclosure [1]. (**g**) No. 6092. Class II equipment with functional earthing/grounding (US). To identify class II equipment (appliances with double insulation and no protective conductor) with functional earthing (grounding) [1]

14.3 EMC Grounding Philosophy

A ground is often thought to be equipotential, meaning there is no voltage difference along a ground conductor or a ground plane. However, this is not true and there is a voltage drop across the $\underline{Z}_g$ [Ω] when current flows along a ground conductor or a ground plane. This voltage drop could lead to *common impedance coupling* (see Sect. 12.1.1). The ground impedance can be written as (Fig. 14.2):

$$\underline{Z}_g = R_g + j\omega L_g \tag{14.1}$$

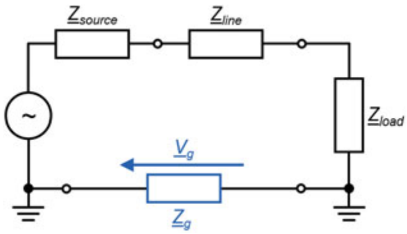


Fig. 14.2 $\underline{Z}_g$ [Ω] is the impedance of the return current signal ground, which is not shown in any circuit schematic but present in every circuit

where:

R_g = ground resistance in [Ω]
L_g = ground inductance in [H]
$\omega = 2\pi f$ = angular frequency of the sinusoidal signal in [rad/sec]

Another point to remember is that the radiated emission due to differential-mode current is proportional to the current loop area A [m^2] (see Sect. 9.9.1). Thus, care must be taken that the current loop area A, formed by the signal current flowing along the signal conductor and through ground back to the signal source, is as small as possible.

The EMC grounding philosophy includes the following points:

- **Ground is a return current path.** Do always consider signal ground as return current path, which closes the signal current loop.
- **Minimize circuit loop area.** With the return current in mind, minimize the current loop area A [m^2], because the radiated emissions of differential-mode current loops are proportional to the current loop area (see Sect. 9.9.1).
- **Consider common impedances.** When multiple signals share a common ground conductor path, common impedance coupling occurs. This coupling can be minimized by minimizing the common impedance $\underline{Z}_g$ [Ω] or by forcing the return currents to flow through different paths.

14.4 Return Current Path on PCB Ground Planes

Current does always take the path of the least impedance $\underline{Z}$ [Ω]. Therefore, in case of a PCB signal trace above a solid ground plane (see Fig. 14.3), we can state the following:

- **At low frequencies** (around $f < 100\,\text{kHz}$), the signal return current path in the *ground plane* flows more or less straight from via 2 back to via 1. This is because at low frequencies, the resistance R_g [Ω] dominates the impedance $\underline{Z}_g = R_g + j\omega L_g$ of the *return current path*, and the lowest resistance from via 2 to via 1 is a direct line.

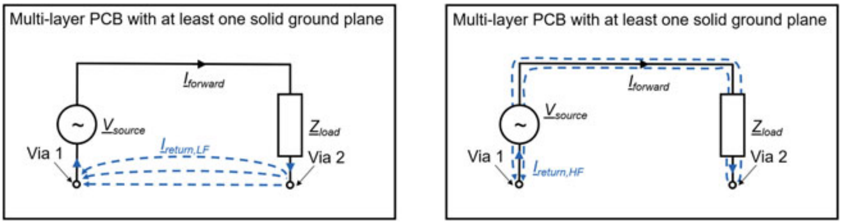


Fig. 14.3 Signal return currents in a solid PCB ground plane. (**a**) At low frequencies (around $f < 100\,\text{kHz}$). (**b**) At high frequencies (around $f > 1\,\text{MHz}$)

- **At high frequencies** (around $f > 1\,\text{MHz}$), the signal return current in the ground plane flows directly below the respective PCB trace. This is because the inductive reactance $X = 2\pi f L_g$ starts to dominate the impedance $\underline{Z}_g = R_g + j\omega L_g$ of the return current path at high frequencies and the inductive reactance is lowest when the return current flows directly underneath the forward current PCB trace.

The points above lead to the conclusion that a solid ground plane should not be split up underneath a high-speed data line because in such a case, the signal return current would have to flow around that gap in the ground plane. This would lead to increased differential current loop area and therefore unintended emissions (see Sect. 9.9.1). Generally speaking, a solid ground plane without any gaps does often lead to the lowest unintended emissions.

Bear in mind that a digital signal—like a clock or data signal—consists of many harmonics where the amplitudes of the high-frequency harmonics primarily depend on the signal's rise- and fall-time (see Sect. 5.2). Given this, the return current path of a digital signal on a PCB is not identical for all its harmonics: the higher the frequency, the closer does the harmonic signal follow the line underneath the forward current path of the digital signal.

14.5 Grounding of Systems

A system in the context of *grounding of systems* means, e.g., a PCBA with multiple circuits placed on the PCB or an electric device which consists of several PCBAs and/or electronic modules and units. We call these circuits, PCBAs, modules, and units in the following just circuits. The grounds of these circuits can be connected in many different ways. Basically, there are three variants [2]:

- **Single-point ground systems.** See Sect. 14.5.1.
- **Multipoint ground systems.** See Sect. 14.5.2.
- **Hybrid ground systems.** See Sect. 14.5.3.

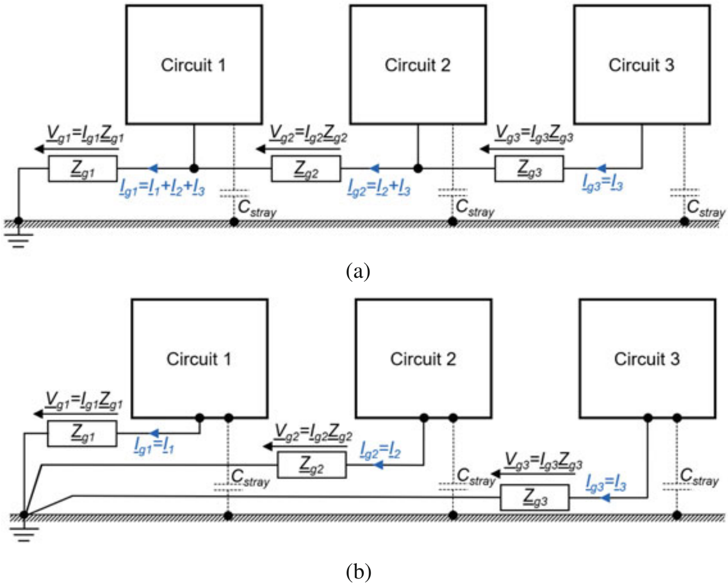


Fig. 14.4 Single-point ground systems. Ideally: $\underline{Z}_{g1} = \underline{Z}_{g2} = \underline{Z}_{g3} = 0\Omega$, $C_{stray} = 0\text{F}$. (**a**) Daisy-chain ground system. (**b**) Star ground system

14.5.1 Single-Point Ground Systems

Single-point ground systems are only useful for low frequencies (<100 kHz). It is nearly impossible to implement a single-point ground system at high frequencies (<1 MHz) because high-frequency currents start to flow through the stray capacitances C_{stray} [F], thus converting a single-ground system into an unintended and, therefore, not optimal multi-ground system.

Figure 14.4 shows two types of single-point ground systems:

- **Daisy-chain ground system.** The *daisy-chain ground system* is a simple way to implement a single-point ground system, but it is not recommended because of the common-impedance coupling.
- **Star ground system.** A *star ground system* is an improved way to implement a single-point ground system than the daisy-chain variant because it reduces the common-impedance noise coupling.

14.5.2 Multipoint Ground Systems

The concept of a *multi-ground system* —like shown in Fig. 14.5—is usually applied for high-frequency systems. At high frequencies (>1 MHz), the ground impedance

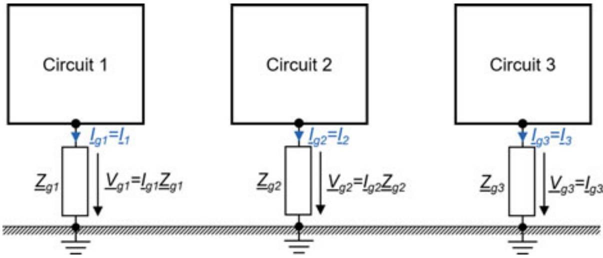


Fig. 14.5 Multipoint ground system. Ideally: $\underline{Z}_{g1} = \underline{Z}_{g2} = \underline{Z}_{g3} = 0\Omega$

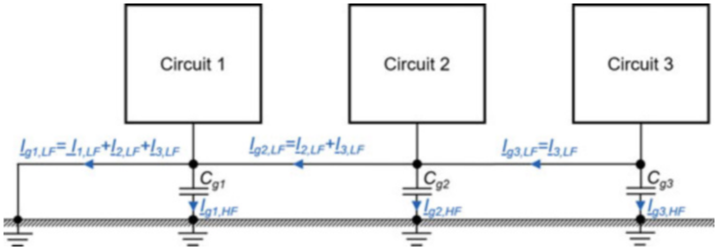


Fig. 14.6 Hybrid ground system. Low-frequency (LF) signals flow to the left to the single-ground point and high-frequency (HF) signals through the capacitors C_{g1}, C_{g2}, and C_{g3}

$\underline{Z}_g = R_g + j\omega L_g$ is primarily dominated by the ground inductance L_g [H]. This means that the ground connections of the circuits to ground should be as short as possible with low impedance (e.g., with multiple vias). For PCBs, a multipoint ground system is best achieved with one or several solid and uninterrupted ground plane(s).

14.5.3 Hybrid Ground Systems

Hybrid ground systems provide different paths to ground for low-frequency currents $\underline{I}_{LF}$ [A] and high-frequency currents $\underline{I}_{HF}$ [A]. Figure 14.6 shows a typical hybrid ground system, which provides a single-point ground system for low-frequency signals and a multipoint ground system for high-frequency signals.

Hybrid grounding can also be applied to cable shields (see Sect. 13.7.4), where one end of the cable shield is connected to ground with low impedance and the other end is connected via a capacitor. A hybrid grounded cable shield could provide reasonable protection against inductive coupling of HF magnetic fields and at the same time prevent LF currents from flowing along the cable shield.

14.6 Ground Loops

A *ground loop* is a current loop formed by ground conductors and ground itself, like shown in Fig. 14.7. Ground loops could lead to the following interference problems:

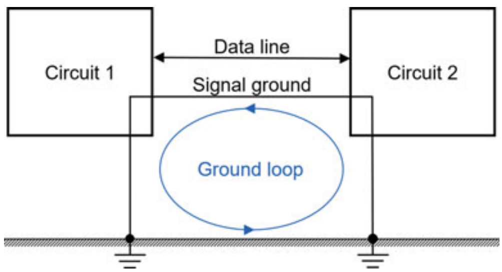
- **Interference caused by ground voltage potential difference.** A voltage potential difference between the ground connection points of two circuits, which are interconnected via a ground conductor, could cause an unintended noise current $\underline{I}_{sg}$ [A] along the signal ground wire between the two circuits. As a consequence, a noise voltage $\underline{V}_n = \underline{I}_{sg} \cdot \underline{Z}_{sg}$ is introduced in the signal connection between the circuits, where $\underline{Z}_{sg}$ [Ω] is the impedance of the signal ground wire of the interconnection between the two circuits.
- **Interference caused by magnetic field coupling to ground loop.** A voltage $\underline{V}_g$ [V] could be induced into the ground loop by a magnetic field (see magnetic field coupling in Sect. 12.1.3 on page 195). $\underline{V}_g$ [V] causes a noise current $\underline{I}_{sg}$ [A] through the signal ground wire, which introduces a noise voltage $\underline{V}_n = \underline{I}_{sg} \cdot \underline{Z}_{sg}$ to the signal interconnection.

Figure 14.8 compares how ground loop interference does affect single-ended (unbalanced) and differential signal (balanced) interfaces. Well-balanced differential interfaces (e.g., LVDS) are more robust against ground loop coupling than single-ended interfaces (e.g., CMOS) because the noise voltage $\underline{V}_n$ [V] affects both signal lines of a differential interface and is canceled out. A list of digital single-ended and differential signal interfaces can be found in Sect. 7.10 on page 90.

In many cases, ground loops do not lead to EMC problems. Meaning, the EMC design engineers should not avoid ground loops at any cost. However, EMC design engineers should be aware of ground loops and how to deal with them if they lead to an EMC issue. Ground loops are primarily a problem for low-frequency applications ($f < 100$ kHz) because the size of ground loops is usually large. Thus, they have a large inductance, which means that they represent a high impedance for HF signals. A typical example of an EMC issue due to a ground loop is the 50/60 Hz hum coupled into audio systems.

If ground loops are a problem, the following are options for minimizing their impact:

Fig. 14.7 Example of a ground loop in case of a single-ended interface



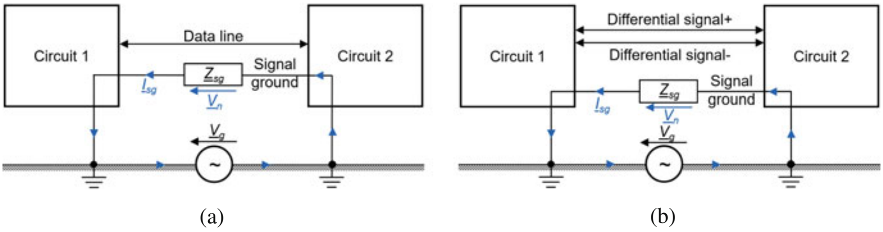


Fig. 14.8 Ground loop noise coupling: single-ended vs. differential interfaces. $\underline{V}_g$ [V] could be caused by a voltage potential difference between the ground connections of circuits 1 and 2, or $\underline{V}_g$ [V] could be induced by a magnetic field. (**a**) Ground loop noise coupling in case of a single-ended (unbalanced) interface. The noise voltage $\underline{V}_n$ [V] does directly interfere with the useful signal. (**b**) Ground loop noise coupling in case of a differential (balanced) interface. $\underline{V}_n$ [V] does add to both differential signal lines likewise, and the noise voltage is canceled out

- **Use balanced transmission lines.** As mentioned above and shown in Fig. 14.8, differential interfaces with a balanced transmission line are very robust against interference caused by ground loops.
- **Applying single-point or hybrid grounding.** Single-point or hybrid ground systems (see Figs. 14.4 and 14.6) help avoid ground current loops caused by voltage potential differences in the ground system. However, they do not help to avoid ground loop currents induced by magnetic fields.
- **Reducing the ground impedance.** Reducing the ground impedance $\underline{Z}_{sg}$ [Ω] in the signal ground conductor—like shown in Fig. 14.8a—reduces the noise voltage $\underline{V}_n$ [V] caused by the unintended ground loop current $\underline{I}_{sg}$ [A] through the signal ground conductor.
- **Breaking the ground current loop:**
  - **Transformers, photocouplers.** If the interface between circuit 1 and 2 in Fig. 14.8 is realized with a transformer or a photocoupler (see Fig. 14.9a,b), the low-frequency ground current path is interrupted. However, HF currents may still flow through the stray capacitance of the transformer or photocoupler.
  - **Common-mode chokes.** If a common-mode choke is added to the interface between circuits 1 and 2 (see Fig. 14.9c), the HF ground loop currents through the signal ground conductor are attenuated. However, at very low frequencies of $f < 10$ kHz and very high frequencies of $f > 1$ GHz, the attenuation of the ground loop currents due to a common-mode choke is limited because of its relatively low inductance L [H] and its stray capacitance C [F].
  - **Fiberglass.** Using a fiberglass communication system between circuits 1 and 2—like shown in Fig. 14.9d—breaks the ground loop for currents of any frequency f [Hz]. However, fiberglass communication systems are costly.

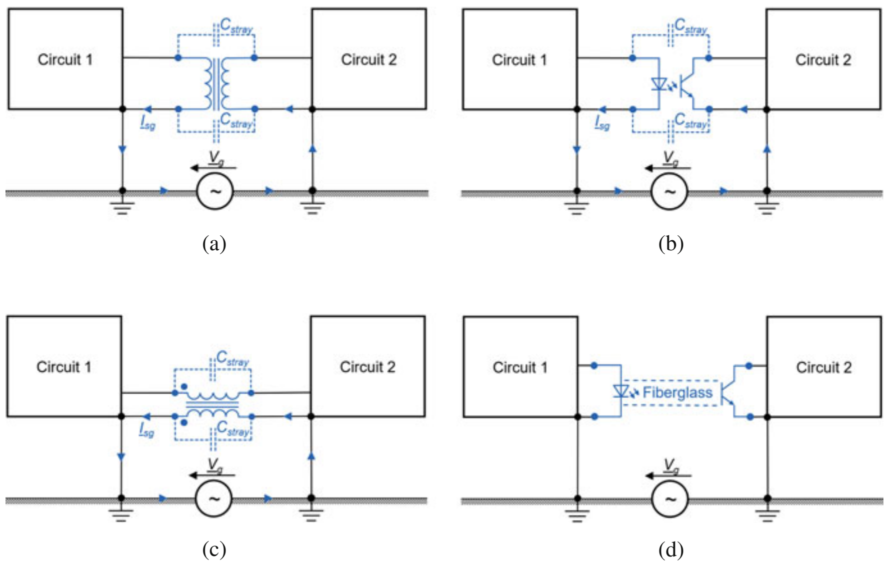


Fig. 14.9 How to break ground loops. (**a**) A transformer can be used to break the ground loop for low-frequency ground loop currents. (**b**) A photocoupler can be used to break the ground loop for low-frequency ground loop currents. (**c**) A common-mode choke can be used to break the ground loop for high-frequency ground loop currents. (**d**) A fiberglass communication system can be used to break the ground loop for noise currents for any frequency f [Hz]

14.7 Summary

- **Types of ground.** There are safety, functional, chassis, and signal grounds.
- **EMC ground philosophy.**
  - Do always consider ground as a return current path.
  - Minimize current loop area between forward and return current path.
  - Consider and minimize common impedance coupling in the ground path.
- **Return current path on solid PCB ground planes**.
  - Signals with $f < 100\,\text{kHz}$: the return current path takes the path of the least resistance and takes the shortest way possible through the ground plane.
  - Signals with $f > 1\,\text{MHz}$: the return current path takes the path of the least inductance and flows back to the source directly underneath the PCB signal trace of the forward current.
- **Grounding of systems**.
  - Single-point ground systems are typically used at LF ($f < 100\,\text{kHz}$).
  - Multipoint ground systems are typically used at HF ($f > 1\,\text{MHz}$).
  - Hybrid ground systems split ground return current paths for LF and HF.

- **Balanced vs. unbalanced transmission lines.** Balanced transmission lines are more robust against ground noise than unbalanced transmission lines.
- **Ground loops.**
  - Ground loop currents are typically caused by:
    - Voltage potential difference in the ground system.
    - Magnetic fields coupling to the ground current loop.
  - Reducing the ground impedance reduces the ground noise voltage caused by ground loop currents.
  - Single-point and hybrid ground systems help reduce ground loop currents.
  - Breaking the ground loop eliminates any ground noise current.

References

1. *Graphical Symbols for Use on Equipment*. International Electrotechnical Commission (IEC). 2002.
2. Henry W. Ott. *Electromagnetic Compatibility Engineering*. John Wiley & Sons Inc., Sept. 11, 2009.

Chapter 15
Filtering

If everyone is thinking alike, then no one is thinking.

—Benjamin Franklin

15.1 Filter Characterization

A linear and time-invariant (LTI) filter can be characterized in the time- and the frequency-domain (Fig. 15.1):

- **Time-domain.** In the time-domain, a filter can be characterized by its *impulse response* $h(t)$ or its *step response* $s(t)$.
- **Frequency-domain.** In the frequency-domain, a filter can be characterized by its *frequency response* $\underline{H}(j\omega)$ (Fourier transform, bode plot).

In this chapter, we focus on filters which are causal *LTI-systems* (except for the non-linear transient filters in Sect. 15.9 and some non-linear digital algorithms in Sect. 15.10). LTI-systems can be completely characterized by one of these functions [4]:

- Impulse response $h(t)$
- Step response $s(t)$
- Transfer function $H(s)$

Causal, *linear*, and *time-invariant* (LTI) means:

- **Causal.** A system is causal if the output depends only on present and past but not on future input values.
- **Linear.** A system is linear if the output $y(t)$ is a linear mapping of the input $x(t)$. If a and b are constants, the output for $a \cdot x(t)$ is $a \cdot y(t)$ and $b \cdot y(t)$ for $b \cdot x(t)$. In addition, for linear systems, you can apply the principle of superposition (the output for $x = a \cdot x_1 + b \cdot x_2$ is $y = a \cdot y_1 + b \cdot y_2$).
- **Time-invariant.** A system is time-invariant when the output $y(t)$ for a given input $x(t)$ does not depend on whether $x(t)$ is applied now or after a time T [sec]. If the output for $x(t)$ is $y(t)$, then the output for $x(t - T)$ is $y(t - T)$.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_15

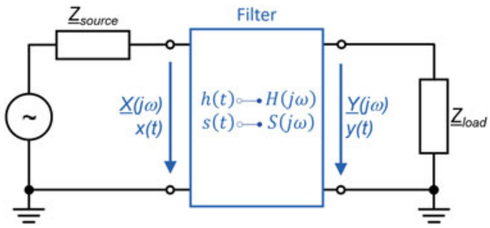


Fig. 15.1 A filter is typically described in the frequency-domain by its frequency response $\underline{H}(j\omega)$ (Bode plot) or in the time-domain its step response $s(t)$

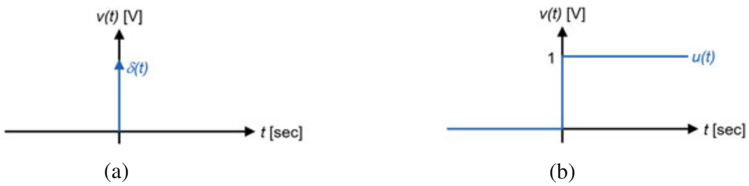


Fig. 15.2 Time-domain characterization functions. (**a**) Impulse Dirac delta function $\delta(t)$. (**b**) Unit step function $u(t)$

Note: real-world components show non-linear effects, e.g., due to current saturation I [A] (inductors, ferrites) or non-ideal high-frequency behavior (inductors, capacitors). Therefore, real-world filters are only linear for certain operating conditions. The non-linearities and other undesired effects of filter components are described in Chap. 11. In this chapter, we assume that the filter components are used in their linear operating region.

15.1.1 Time-Domain: Step Response

The goal of a filter characterization in the time-domain is to determine signal distortions and the time delay caused by the filter (Fig. 15.2).

If the input signal $x(t)$ of an LTI-system is the *Dirac delta* $\delta(t)$ function:

$$\delta(t) = \begin{cases} +\infty & t = 0 \\ 0 & \text{otherwise,} \end{cases} \tag{15.1}$$

then the output is defined as the *impulse response* $h(t)$. The impulse response characterizes an LTI-system entirely, and for any given input signal $x(t)$ the output $y(t)$ can be calculated as [4]:

$$y(t) = h(t) * x(t) = \int_{-\infty}^{t} h(t-\tau)x(\tau)d\tau \tag{15.2}$$

where

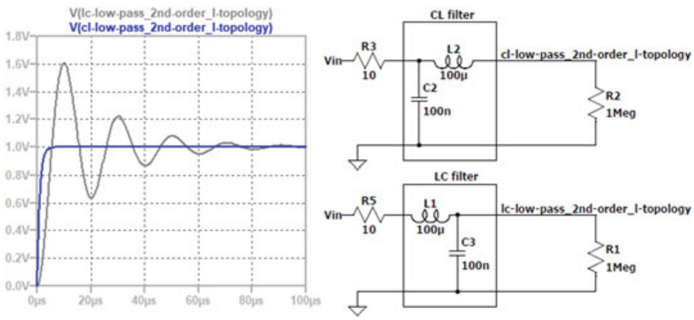


Fig. 15.3 Step response simulation of two 2nd-order low-pass filters [1]

$y(t)$ = the output signal from an LTI-system in [V]
$h(t)$ = the impulse response of the LTI-system in [V]
$x(t)$ = the input signal to an LTI-system in [V]
$*$ = convolution, a mathematical operation of two functions

In practice, the *step response* $s(t)$ (and not the impulse response $h(t)$) is usually used to characterize a filter [4]:

$$u(t) = \begin{cases} 1 & t \geq 0 \\ 0 & t < 0. \end{cases} \tag{15.3}$$

If the input is the *unit step*: $x(t) = u(t)$, the output $y(t)$ is called step response $s(t)$. The step response $s(t)$ can be calculated by integrating the impulse response $h(t)$ and the impulse response $h(t)$ by taking the derivative of the step response $s(t)$ [4]:

$$s(t) = \int_{-\infty}^{t} h(\tau) d\tau \tag{15.4}$$

$$h(t) = \frac{d}{dt} s(t) \tag{15.5}$$

Figure 15.3 shows the step response of two different low-pass filters, where:

- Step signal amplitude $u(t) = 1$ V for $t > 0$ sec
- Low source impedance $\underline{Z}_{source} = R_{S1} = R_{S2} = 10\,\Omega$
- High load impedance $\underline{Z}_{load} = R_{L1} = R_{L2} = 1\,\text{M}\Omega$
- Filter type = 2nd order passive low-pass filter
- Filter topology = L-topology

The step responses in Fig. 15.3 show that even passive low-pass filters can cause a significant amount of ringing and time delay. Therefore, it is always necessary to also check the time-domain behavior of an EMI-filter, not just in the frequency-domain behavior.

15.1.2 Frequency-Domain: Frequency Response

The *frequency response* $\underline{H}(j\omega)$ (*Fourier transform*) is the transfer function $\underline{H}(s)$ (*Laplace transform*, $s = \sigma + j\omega$) of a filter for a sinusoidal input signal, meaning $s = j\omega$. The frequency response $\underline{H}(j\omega)$ describes the attenuation in [dB] and the phase shift in [rad] or [°] as a function of frequency $\omega = 2\pi f$.

The Fourier transform of the filter's impulse response $h(t)$ is the frequency response $\underline{H}(j\omega)$ of the filter [4]:

$$h(t) \circ\!\!-\!\!\bullet \underline{H}(j\omega) = \int_{-\infty}^{\infty} h(t)e^{-j\omega t}dt \tag{15.6}$$

The frequency response characterizes a filter in the frequency-domain. For any given input signal $\underline{X}(j\omega)$, the output signal $\underline{Y}(j\omega)$ can be calculated as [4]:

$$\underline{Y}(j\omega) = \underline{H}(j\omega) \cdot \underline{X}(j\omega) \tag{15.7}$$

where

$\underline{Y}(j\omega) =$ the output signal of the filter in [V]
$\underline{H}(j\omega) =$ the frequency response of the filter in [V]
$\underline{X}(j\omega) =$ the input signal of the filter in [V]
$\omega = 2\pi f =$ angular frequency of the sinusoidal input signal in [rad/sec]

The frequency response $\underline{H}(j\omega)$ can also be calculated by using the Fourier transform of the step response $\underline{S}(j\omega)$:

$$s(t) \circ\!\!-\!\!\bullet \underline{S}(j\omega) \tag{15.8}$$

$$\underline{H}(j\omega) = j\omega \cdot \underline{S}(j\omega) \tag{15.9}$$

In practice, the frequency response $\underline{H}(j\omega)$ is shown in the form of a *Bode plot*. Figure 15.4 presents the Bode plot of the identical filters of Fig. 15.3, which shows us a couple of interesting points:

- **Ringing and resonance.** It can be seen that the ringing in the time-domain matches the resonance in the frequency-domain at roughly $f_r = 50\,\text{kHz}$.
- **Order vs. attenuation and phase shift.** The number of reactive components gives the order n of an analog filter; in other words, the sum of inductors L [H] and capacitors C [F] of the filter is equal to the filter order n. Because the filters are of second-order ($n = 2$), the attenuation after the cut-off frequency is $n \cdot 20\,\text{dB} = 40\,\text{dB}$ and the maximum phase shift is $n \cdot -90\,\text{deg} = -180\,\text{deg}$.
- **Source and load impedance vs. filter performance.** The filter performance (attenuation, phase shift) depends on the filter components themselves and the source and load impedances. Have a look at the CL low-pass filter in Fig. 15.4, where the source impedance $R_{S2} = 10\,\Omega$ builds a low-pass RC-

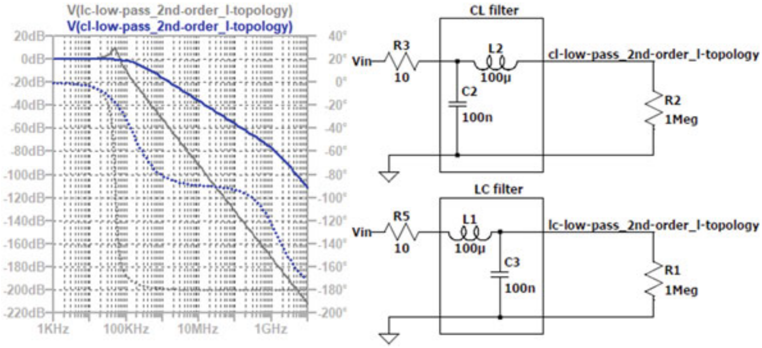


Fig. 15.4 Frequency response simulation of two second-order low-pass filters [1]

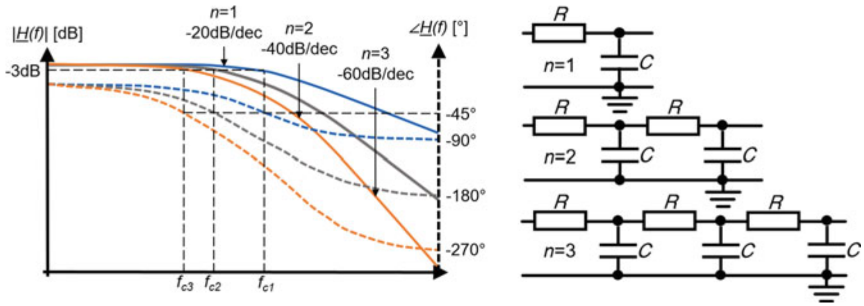


Fig. 15.5 Bode plot of ideal RC low-pass filters of order $n = 1$, $n = 2$ and $n = 3$

filter with the first filter component C_2 [F] which has a cut-off frequency of $f_g = 1/(2\pi R_{S2} C_2) = 160\,\text{kHz}$ and the load impedance $R_{L2} = 1\,\text{M}\Omega$ builds a low-pass RL-filter with the filter component L_2 [H] and a cut-off frequency of $f_g = R_{L2}/(2\pi L_2) = 1.6\,\text{GHz}$.

15.2 Low-Pass Filters

Low-pass filters are the most common filters in the world of EMI and EMC. They reject undesired HF energy above a desired cut-off frequency f_c [Hz]. Figure 15.5 shows the frequency response of ideal low-pass filters (skin effect, stray capacitance, and other non-linearities are neglected). At the cut-off frequency f_c [Hz], the insertion loss is 3 dB and the phase shift is -45 deg. The attenuation for frequencies $f > f_c$ is $n \cdot 20$ dB and the maximum phase shift of a low-pass filter is $n \cdot -90$ deg, where n is the filter order.

In most cases, EMI-filters against radiated or conducted high-frequency (HF) interference are analog passive low-pass filters with resistors, capacitors, inductors, ferrites, or common-mode chokes as filter components. The filter order n is usually not higher than three (maximum phase shift $= n \cdot 90\,\text{deg} = 270\,\text{deg}$) because of the signal distortion and because filters of order $n \geq 4$ could easily lead to instability due to the large phase shift of 360 deg and more.

The filter performance (e.g., cut-off frequency f_c [Hz]) depends not only on the filter components themselves but also on the source impedance $\underline{Z}_S$ [Ω] and the load impedance $\underline{Z}_L$ [Ω]. Therefore, the input and output impedances of the filter should be matched in the useful frequency range and mismatched in the EMI noise frequency range (an overview is given in Fig. 15.24).

It is highly recommended to simulate EMI analog circuit board filters with the real-world source and load impedances and the non-ideal SPICE models of the filter elements. The SPICE simulations of the low-pass filters in Fig. 15.6 show that it is important to consider the non-ideal filter components (not the ideal components) because the non-ideal behavior of capacitors, inductors, and ferrite beads influence the filter performance significantly at high frequencies.

15.3 High-Pass Filters

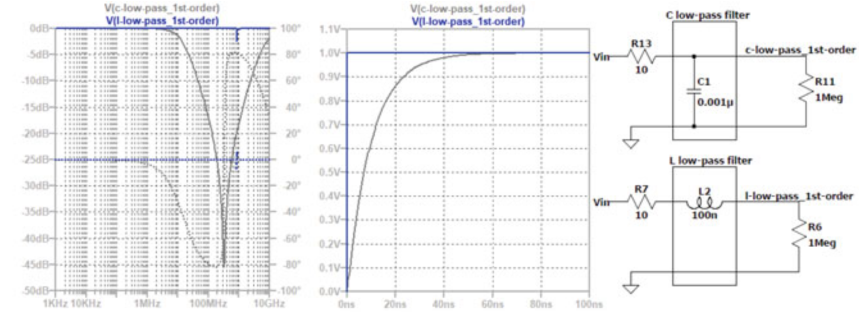
High-pass filters do attenuate low-frequency (LF) signals and let high-frequency (HF) signals pass without attenuating them. Figure 15.7 shows the frequency response of ideal high-pass filters. In the field of EMC, high-pass filters are not very common, and we do not go into further detail about high-pass filters at this point.

15.4 Band-Pass Filters

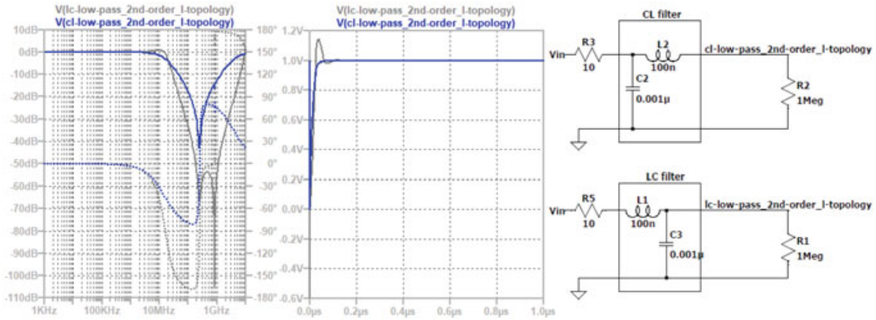
Band-pass filters do only pass signals of the desired frequency range with little to no insertion loss. Figure 15.8 shows the frequency response of an ideal band-pass filter. However, band-pass filters are very seldom used in the field of EMC and are not discussed in further detail here.

15.5 Band-Stop Filters

Band-stop filters do attenuate only signals within the desired frequency range and pass the others. Figure 15.9 shows the Bode plot of a passive band-stop filter with an ideal. Band-stop filters are not very common EMI-filters and therefore not further discussed here.



(a)



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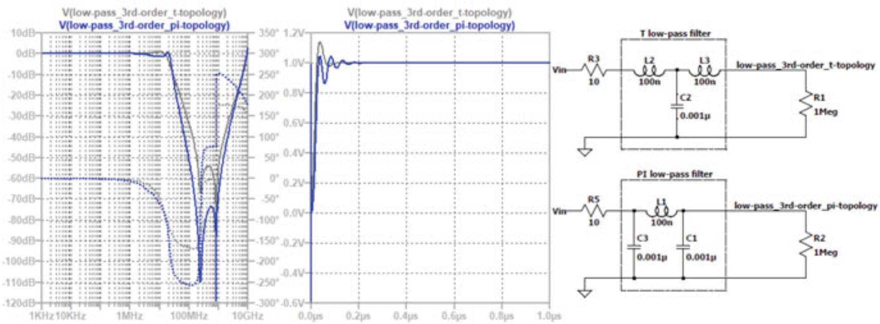

(c)

Fig. 15.6 Passive low-pass filter SPICE simulations with $\underline{Z}_S = 10\,\Omega$ and $\underline{Z}_L = 1\,\text{M}\Omega$. Tool = LTspice [1]. L = Würth 744766001 WE-GF, C = Würth 885012008026 NP0. (**a**) 1st order low-pass filter. (**b**) Second-order low-pass filter. (**c**) 3rd order low-pass filter

15.6 Active and Passive Filters

The filters in the previous sections were all *passive filters*. If operational amplifiers or transistors are part of a filter, it is called an *active filter*. In most cases, EMI-filters

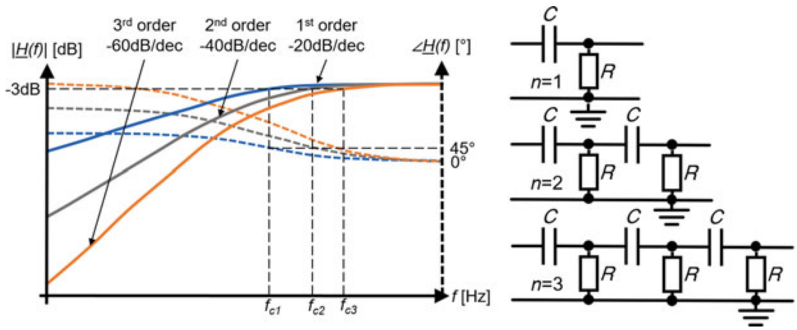

Fig. 15.7 Bode plot of ideal RC high-pass filters of order $n = 1$, $n = 2$, and $n = 3$

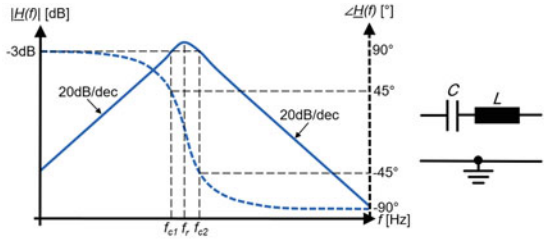

Fig. 15.8 Bode plot of an ideal LC-band-pass filter. f_r [Hz] denotes the resonance frequency and f_{c1} [Hz] and f_{c2} [Hz] the cut-off frequencies

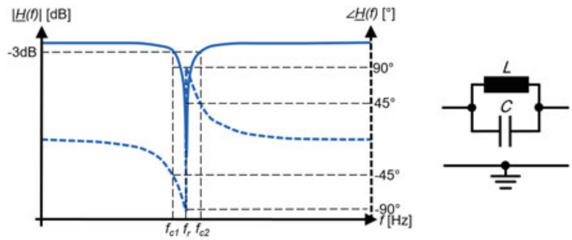

Fig. 15.9 Frequency response of an ideal passive LC band-stop filter. f_r [Hz] denotes the resonance frequency and f_{c1} [Hz] and f_{c2} [Hz] the cut-off frequencies

Table 15.1 Active vs. passive filters: advantages and disadvantages

	Active filter	Passive filter
Advantages	- More design freedom - Steeper pass-band to stop-band transition - Gain > 0 dB possible	- Simple to design (stable) - Small space requirements - Cheap
Disadvantages	- More complex to design (offset voltages, non-linearities, power supply) - Expensive (operational amplifiers) - Possible instability (oscillation)	- Design limitations (gain, pass-band to stop-band transition) - Limited pass-band to stop-band transition - No gain > 0 dB possible

are passive filters because the advantages of active filters mentioned in Table 15.1 are often not needed for EMI-filter applications.

Figure 15.10 compares two second-order passive and active low-pass filters. Although the active filter shows a narrower transition between the pass-band to the stop-band, the attenuation of the active filter at frequencies of $f > 100$ MHz is not

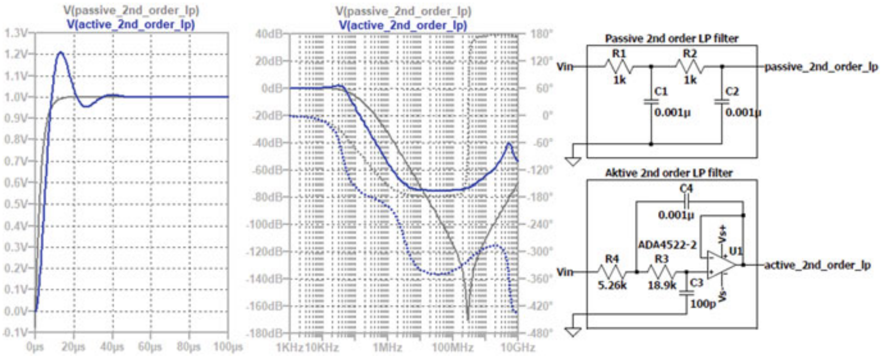

Fig. 15.10 Step and frequency response comparison of an active and a passive low-pass filter with pass-band up to $f_c = 60\,\text{kHz}$ and order $n = 2$ [1]

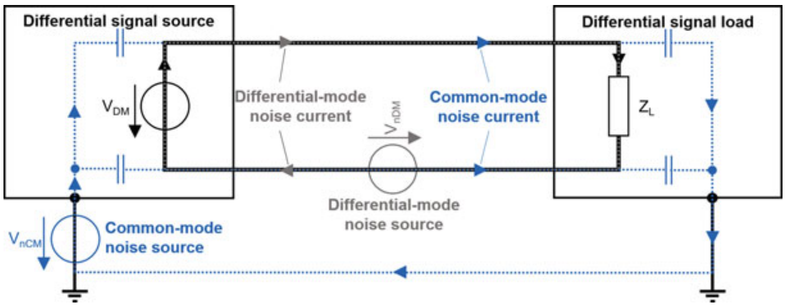

Fig. 15.11 Differential-mode vs. common-mode noise currents

as good as for the passive filter and the active filter shows significant ringing in the time-domain. This example should just illustrate that an active filter is not always better than a passive filter; it all depends on the application's requirements.

15.7 Differential- and Common-Mode Filters

Differential-mode noise and common-mode noise are in detail explained in Sects. 12.3 and 12.4 and compared in Fig. 15.11. The filter components and topology for differential- and common-mode filters are not identical. Therefore, it is necessary to know if you deal with differential-mode or common-mode noise or both.

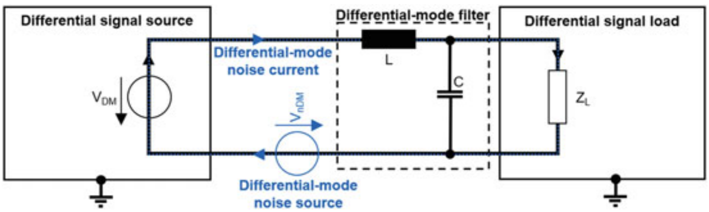

Fig. 15.12 Example of a differential-mode *LC*-filter

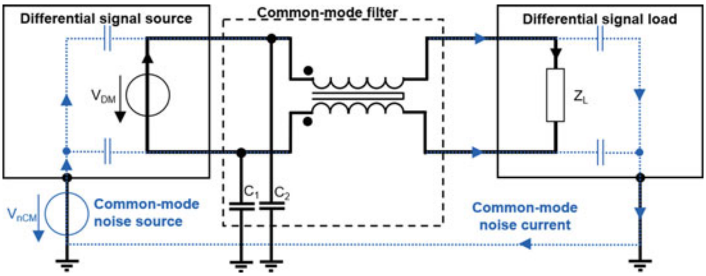

Fig. 15.13 Common-mode filter with Y-capacitors and a common-mode choke

15.7.1 *Differential-Mode Filters*

Differential-mode noise filters are filters between a power or signal line and its return current line (see Fig. 15.12). Typical differential-mode noise filter components are:

- **X-Capacitors.** See Fig. 15.14 and Sect. 11.3 on page 157.
- **Inductors.** See Sect. 11.4 on page 164.
- **PCB mount ferrite beads.** See Sect. 11.5.2 on page 171.

15.7.2 *Common-Mode Filters*

Common-mode noise filters have the purpose to attenuate common-mode noise. An example of a common-mode filter is given in Fig. 15.13. Typical common-mode noise filter components are:

- **Y-capacitors.** See Fig. 15.14 and Sect. 11.3 on page 157.
- **Common-mode chokes.** See Sect. 11.6 on page 173.
- **Cable mount ferrite beads.** See Sect. 11.5.1 on page 169.

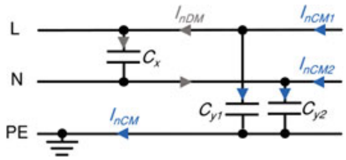

Fig. 15.14 X-capacitors vs. Y-capacitors

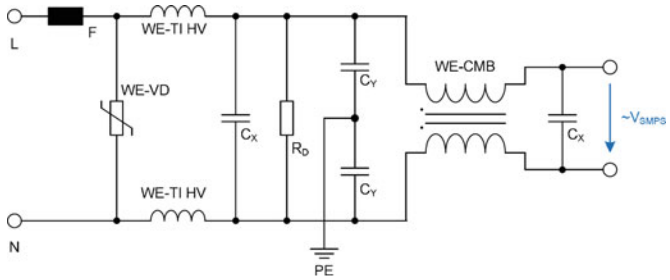

Fig. 15.15 Example of an AC mains filter. Courtesy of Würth Elektronik GmbH

15.8 Mains Supply Filters

Special care must be taken regarding the safety requirements of filter components of (public) mains supply filters (Fig. 15.15). This is especially true for the X- and Y-capacitors:

- **X-capacitors.** X-capacitors are capacitors between line (L) and neutral (N) of the AC mains. A failure (short circuit of the X-capacitor) could result in fire.
- **Y-capacitors.** Y-capacitors are capacitors between line (L) to ground and between neutral (N) to ground of the AC mains. The capacitance of a Y-capacitor must be low enough so that the maximum allowed leakage current through the Y-capacitor is below the required limit!

The classification of X- and Y-capacitors according to IEC 60384-14 [3] is shown in Table 11.1 on page 158.

The applicable safety standards of a product must be considered during the design of a mains supply filter. Examples are:

- **IEC 61010-1.** Safety requirements for electrical equipment for measurement, control, and laboratory use—Part 1: General requirements [5].
- **IEC 62368-1.** Audio/video, information and communication technology equipment—Part 1: Safety requirements [2].
- **IEC 62477-1.** Safety requirements for power electronic converter systems and equipment—Part 1: General [6].

15.9 Transient Suppression Filters

A distinction is made between these three types of high-voltage transients (Table 15.2):

- **Electrostatic discharge (ESD).** Low energy pulses, very short rise-time.
- **Electrical fast transients (EFT).** Bursts of low energy pulses, short rise-time.
- **Lightning surge pulses (surge).** High energy pulses, medium rise-time.

The rise-time and the pulse energy are the most important parameters of voltage transients. Therefore, the pulse forms and energy are presented in Figs. 15.16 (ESD), 15.17 (EFT), and 15.18 (surge). The pulse energies were calculated by considering the maximum tolerances of the test equipment (timing, test voltage). The maximum pulse energy for ESD was calculated by neglecting the internal losses and by just considering the energy E [J] stored in the capacitor of the ESD generator:

$$E_{ESD} = \frac{1}{2}CV_p^2 = \frac{150\,\text{pF}}{2}V_p^2 \tag{15.10}$$

Table 15.2 Overview of high-voltage transient EMC testing parameters

Transient	Open load peak pulse voltage range V_p	Source impedance R_s	Rise-time t_r	Pulse width t_{pw}	IEC standard
ESD	±2–8 kV, contact ±2–15 kV, air	330 Ω	0.7–1 ns current pulse	60 ns, current pulse	61000-4-2
EFT (bursts)	±0.5–4 kV, power ports ±0.25–2 kV, signal ports	50 Ω	5ns, voltage pulse @ 50 Ω load	50 ns, voltage pules @ 50 Ω load	61000-4-4
Surge	±0.5–2 kV, line-to-line ±0.5–4 kV, line-to-ground	2 Ω, line-to-line 12 Ω, line-go-ground 42 Ω, others-to-ground	1.2 μs or 10 μs, voltage pulse @ open circuit	50 μs or 700 μs, voltage @ open circuit	61000-4-5

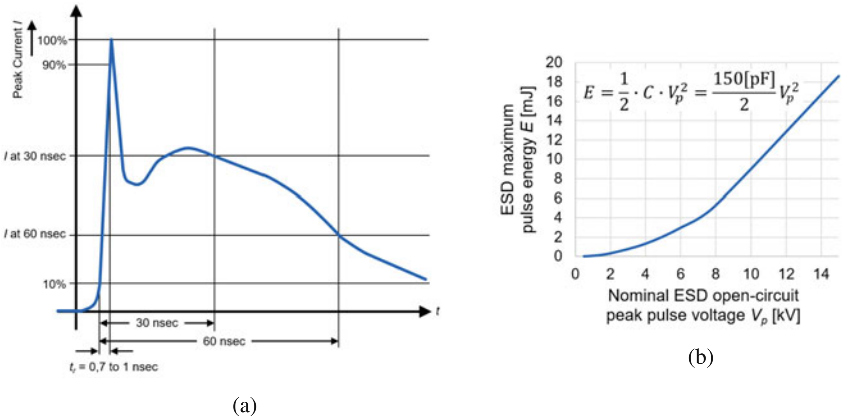

Fig. 15.16 ESD transient pulse. (**a**) ESD current pulse form. (**b**) Approximate maximum ESD-pulse energy

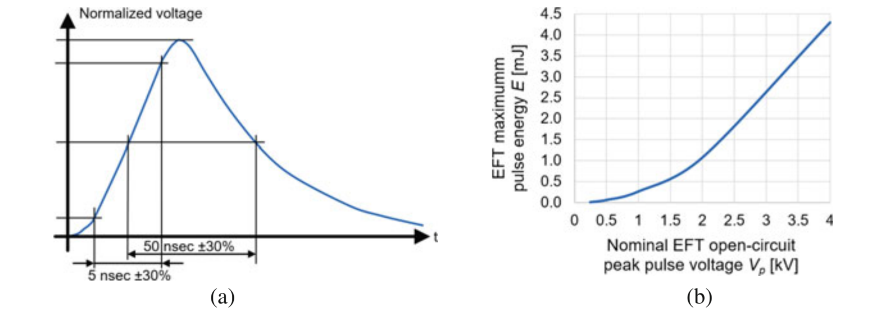

Fig. 15.17 Single EFT transient pulse. (**a**) Single EFT-pulse form (open-circuit voltage). (**b**) Approximate maximum single EFT-pulse energy

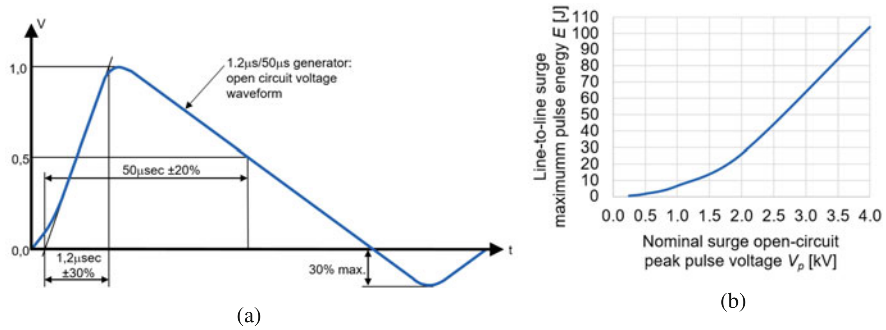

Fig. 15.18 Surge transient pulse. (**a**) Surge-pulse form (open-circuit voltage). (**b**) Approximate maximum 1.2/50 µs surge-pulse energy (line-to-line)

where

$C = 150\,\text{pF}$ = capacitance of the ESD generator's energy storage capacitor
V_p = ESD test voltage in [V]

The maximum EFT- and surge-pulse energy E [J] was calculated by approximating the pulse energy with the following formula [8]:

$$E_{EFT,surge} = \left(\frac{1}{3} \cdot \frac{V_p^2}{R} \cdot t_1\right) + \left(\frac{V_p^2 \cdot \tau}{-2 \cdot R}\right) \cdot \left(e^{-\frac{2(t_3 - t_1)}{\tau}} - 1\right) \tag{15.11}$$

where

V_p = EFT/surge peak test voltage in [V]
R = load resistance of the EFT/surge test generator in [Ω]
t_1 = time to peak voltage in [sec]
t_2 = time until voltage has fallen off to 50 % of peak voltage in [sec]
t_3 = time to negligible voltage in [sec]
$\tau = -(t_2 - t_1)/\ln(0.5)$ = exponential rate of decay in [sec]

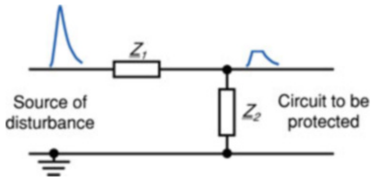

Fig. 15.19 High-voltage transient suppressor filter stage

As ESD- and EFT-pulses show similar rise-times [nsec] and pulse energies [mJ], thus, they can usually be filtered with the same filter components. However, dealing with surge-pulses means dealing with potentially more than 1000 times higher energy compared to ESD or EFT transients. Thus, surge filters do usually need components with high energy absorption capability.

A transient filter stage consists typically of two components (see Fig. 15.19):

- **Series impedance $\underline{Z}_1$**. The series impedance limits the transient current through the shunt element and the circuit. Typical series protection components are: *resistors*, *ferrites*, or the series resistance and inductance of the *conductor* itself.
- **Shunt impedance $\underline{Z}_2$**. The shunt element limits the transient voltage across the circuit. Typical shunt protection components are clamping devices like *TVS diodes* (see Sect. 11.8.2) and *varistors* (see Sect. 11.8.1) or crowbar devices like *thyristors* or *gas discharge tubes* (see Sect. 11.9), or simply *capacitors* (see Sect. 11.3).

In the following, we focus on the selection of an appropriate shunt protection element $\underline{Z}_2$ [Ω]. The following points must be considered when choosing a voltage protection shunt element:

- $\mathbf{V_{operation} < V_{standoff}}$. The maximum operating voltage must be smaller than the standoff voltage of the shunt element. Up to the standoff voltage (reverse working voltage), the shunt element does not conduct any significant amount.
- $\mathbf{V_{clamp} < V_{damage}}$, $\mathbf{V_{trigger} < V_{damage}}$. With the maximum clamping voltage or the trigger voltage, the maximum voltage across the shunt element is meant. This voltage must be lower than the voltage at which the circuit to be protected experiences a defect or malfunction.
- **Energy absorption:** $\mathbf{E_{pulse} < E_{absorption}}$. The protective elements must absorb the energy (voltage, current) of the expected transient pulse without any damage.
- **Signal integrity:** $\mathbf{C_{shunt} < C_{max}}$. The protective elements must not interfere with the useful signal (e.g., due to high capacitance or leakage current) and the signal integrity must be guaranteed.
- **Lifetime:** $\mathbf{n_{expected} < n_{max}}$. The protective elements must guarantee to withstand the minimum number of expected transient events *n* during the lifetime, and the protective devices' degradation must be tolerable.

15.10 Digital Filters

Up to this point, all presented filters in this chapter were *analog filters*. This section is dedicated to *digital filters*, which have a couple of advantages (Table 15.3).

Here are some fundamental points about digital filters (Fig. 15.20):

- **Nyquist-Shannon sampling theorem**.[1] In order to convert an analog signal $x(t)$ with the bandwidth B [Hz] (highest frequency in the signal) to a digital signal $x[n]$, a minimum *sampling frequency* of $f_s > (2 \cdot B)$ [Hz] is required.
- **Aliasing.** Aliasing happens when the Nyquist-Shannon sampling theorem is violated. In order to prevent aliasing, an analog *anti-aliasing low-pass filter* is placed in front of every ADC. This low-pass filter typically has a cut-off frequency f_c [Hz] that is much lower than half of the sampling frequency f_s [Hz]: $f_c \ll f_s/2$.
- **FIR-filters.** *FIR-filters* have finite impulse responses (FIR) $h[n]$ and do not have feedback coefficients. Linear moving-average filters are FIR filters.
- **IIR-filter.** *IIR-filters* have an infinite impulse response (IIR) $h[n]$, use feedback coefficients, and provide more design freedom than FIR-filters. Digital filters with Butterworth, Bessel, or Chebyshev behavior require a digital IIR-filter.

Table 15.3 Digital vs. analog filters

	Analog filters	Digital filters
Advantages	- High-speed, wide bandwidth - Flexible placing within the system (e.g. at the entrypoint of a cable)	- Less hardware, lower costs - Versatile and flexible adaption of filter behavior
Disadvantages	- Hardware nacessary, higher costs - Not flexible regarding changes	- Bandwidth depends on sampling rate (Nyquist-Shannon sampling theorem) - Physical location is inside microcontroller

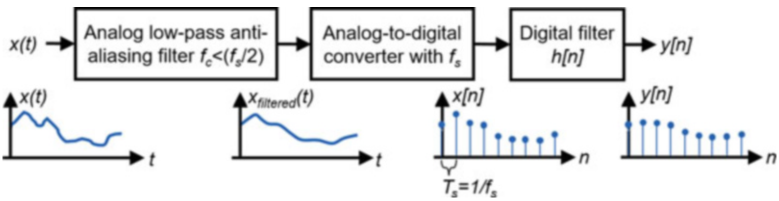

Fig. 15.20 Digital filter with ADC. f_c = cut-off frequency of analog anti-aliasing low-pass filter $f_s = 1/T_s$ = sampling-frequency of the ADC

[1] The original Nyquist-Shannon sampling theorem [7]: If a function $f(t)$ contains no frequencies higher than B [Hz], $f(t)$ is completely determined by giving its ordinates at a series of points spaced $1/(2B)$ seconds apart.

Even though the usage of digital filters in EMC is limited, they can help improve robustness, especially when it comes to voltage transient immunity. Here are some typical examples:

- **Debounce filter.** Analog or digital inputs are read multiple times and checked for unexpected state changes. For example, it could be determined that an input state is only valid when it is stable over a certain period (debounce time).
- **Analog sensor sanity check.** When reading an analog sensor input, e.g., a pressure or a flow rate, the sensor value could be checked if it is within a reasonable range. If not, the sensor value could either be reread, or the system could throw a warning.
- **Input states sanity check.** In the case of multiple sensors (e.g., position sensors along a linear drive with a slider), a sanity check could be performed (e.g., it is only possible that one single position sensor along the linear drive detects the slider). In the event of an implausible result, a new status query could be triggered, or the system could enter a safe state.
- **Median filter.** Median filters are non-linear filters that provide a reliable way to get rid of short spikes and dips in an input signal. The filter takes an array of N values, sorts it in ascending order, and returns the value in the middle of the sorted array. Figure 15.21 shows the conceptual workflow.
- **Spike-remover.** A spike remover detects a spike in an array of data and replaces the spike data with valid data points. Figure 15.22 shows the principle of a single data point spike-spike remover.

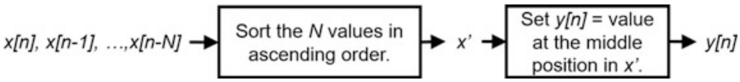

Fig. 15.21 Concept of a median filter

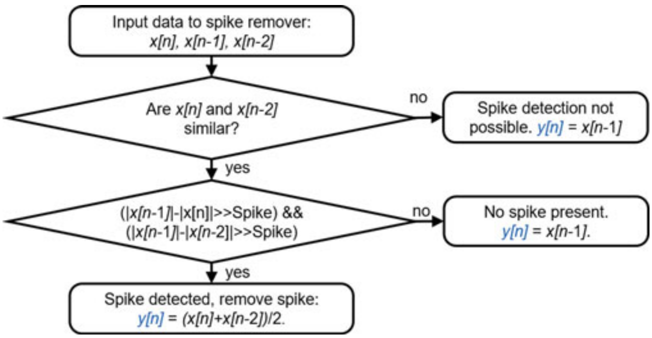

Fig. 15.22 Concept of a spike-remover

15.11 Summary

- **Filter characterization.** EMI-filters should always be characterized by their step response (time-domain) and their frequency response (frequency-domain) (Fig. 15.23).
  - **Time-domain.** Determine distortion and time delay caused by the EMI-filter.
  - **Frequency-domain.** Determine attenuate [dB] and phase shift of the EMI-filter.

- **High-frequency filters.** Filters against radiated and conducted HF emissions and immunity are in most cases passive low-pass filters like shown in Fig. 15.24.

- **Transient suppression filters.** Typically non-linear clamping devices (TVS diodes, varistors) or capacitors are used as protective shunt devices. Figure 15.25 summarizes the most important points when selecting a high-voltage transient suppressor device.

- **Digital filters.** Filters implemented in software and firmware are helpful to improve EMC immunity. Examples are debounce filters, input state sanity checks, sensor value sanity checks, spike-remove filters, and median filters (Fig. 15.26).

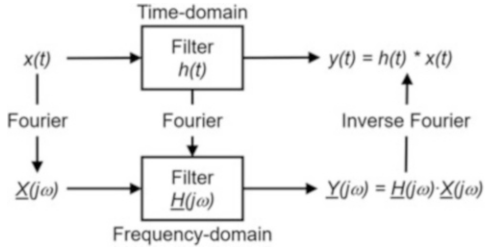

Fig. 15.23 Analog filters: Time-domain vs. frequency-domain

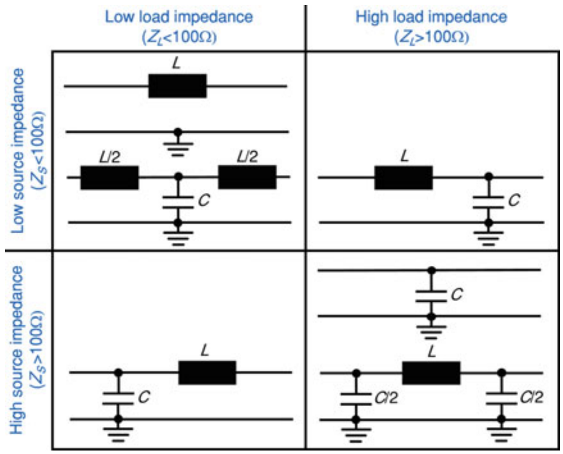

Fig. 15.24 EMI low-pass filter overview

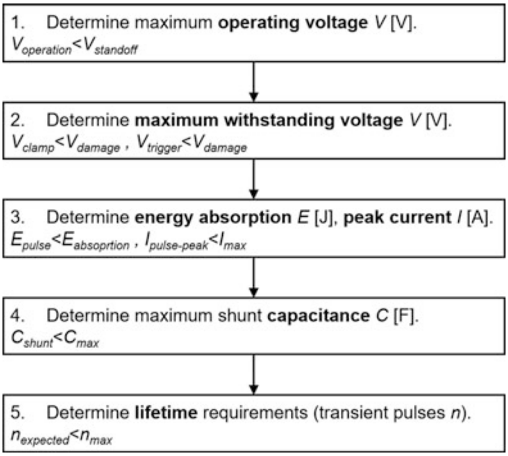

Fig. 15.25 Selecting a protective shunt device for high-voltage transient suppression

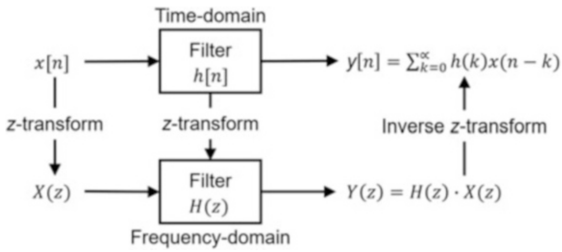

Fig. 15.26 Digital filters: time-domain vs. frequency-domain

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Chapter 16
EMC Design Guidelines

Everything should be made as simple as possible, but no simpler.

—Albert Einstein

16.1 Most Common EMC Issues in Practice

The guidelines in this chapter are biased by the author's own experience, gathered during many years of electronics development, firmware programming, and systems engineering. This should be mentioned here as a kind of a disclaimer and that you, the reader, can better understand how to weigh and classify the guidelines presented here.

From the author's own experience, these are the most common EMC issues during product development:

- **Radiated emissions.** Many products regularly fail radiated emissions EMC testing from 30 MHz to 6 GHz according to CISPR 32 or CISPR 11. Most issues typically occur in the frequency range <1 GHz, where the causes of radiation are usually cables or units within the system which are not grounded properly. If the emission limits for $f > 1$ GHz cannot be met, the unintentional radiator can most probably be found on a PCB (clock lines, single-ended high-speed data lines, small PCB-structures).
- **Radiated immunity.** Radiated immunity for most products requires testing according to IEC 61000-4-3. Usually, sensors are the most sensitive elements in a product. This is why sensors require special attention during the design and EMC immunity testing. Care must be taken that all signal lines to and from a sensor are filtered with the maximum allowed attenuation. However, every filter causes distortions and time delays (phase shift in the frequency-domain). Therefore, ensure that the distortions and time delays are below the allowed maximum. In addition, low-impedance grounding of any sensitive circuit is of utmost importance and the cabling and PCB-structures should be kept as short and small as possible.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7_16

- **Electrostatic discharge.** Grounding, grounding, grounding. ESD testing is performed according to IEC 61000-4-2. Every electrically floating part of a product increases the chance of failing the ESD-test because it increases the chance that undesired currents—caused by ESD-pulses—flow along wires and PCB signal lines and eventually lead to damaged integrated circuits or disturbed signals.

16.2 Guideline # 1: Never Route Signals Over Split Reference Planes

Goal = reduced radiated emissions.
Do not route signals over split return signal reference planes (GND, power planes)! Never! This leads to unnecessary large current loops (as the current return cannot flow directly underneath the forward current, Fig. 16.1) and large current loops tend to lead to high radiated emission values.
Rule of thumb: Whenever in doubt, do not split return current reference planes (GND, power planes). Go with a solid-filled reference plane instead. There must be a good reason for splitting planes! There should always be at least one solid reference plane closely adjacent to high-frequency signals ($f > 1$ MHz).

16.3 Guideline # 2: Always Consider the Return Current

Goal = reduced radiated emissions and common-impedance coupling.
Always consider the return current! Always! And with the return current in mind, minimize the common return current paths of high-current circuits (e.g., motors) and sensitive circuits (e.g., unamplified sensor signals) because these common paths lead to common-impedance coupling (see Sect. 12.1.1). In addition, try to minimize the loop area A [m^2] between the forward and return current of high-frequency signals because these loops may lead to unintended radiated emissions (see Sect. 9.9.1).
Hint: Assuming a high-frequency signal s_1 ($f > 1$ MHz) along a PCB trace on layer 1 and a reference plane (e.g., GND or power plane) on layer 2, the return current of the high-frequency signal s_1 on the reference plane prefers to flow directly under the trace of s_1 because the high-frequency return current always takes the path of the least inductance (see Sect. 14.4).

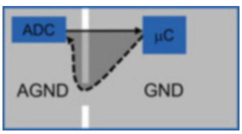

Fig. 16.1 Signal and its return current path in case of split ground planes

16.4 Guideline # 3: Decoupling—Use Low-Inductance Capacitors AND Planes

Goal = reduced radiated emissions and common-impedance coupling.
Decoupling is important! (Fig. 16.2) Always consider decoupling!
Hint: Decoupling depends on the number of PCB layers and stackup:

- **Two- and single-layer PCBs.** For single- or two-layer PCB designs, place the decoupling capacitors as close as possible to every power supply pin of every chip of the PCB. This refers to *local decoupling*.
- **Multi-layer PCBs.** In the case of multi-layer PCB designs with closely spaced GND and power planes (<0.25 mm, <10 mils), the decoupling capacitor location is not as critical as for single- or two-layer designs because the closely spaced planes act as an efficient decoupling capacitor for high-frequencies (>1 MHz) and the actual decoupling capacitor acts as a *global decoupling* capacitance. However, more important than the location of the decoupling capacitor is that the power supply and GND pins of the integrated circuits are connected with low inductance (multiple vias) to the respective planes.

16.5 Guideline # 4: Use Ground Planes on PCB for Shielding

Goal = reduced interference on circuit board.
Use solid-filled reference planes (e.g., GND or power supply planes) on a PCB to separate noise signals (e.g., motor signals) from sensitive signals (e.g., unamplified sensor signals). The reference plane will act as a shield and will lower electromagnetic interference (EMI). However, be aware that such a copper shield will

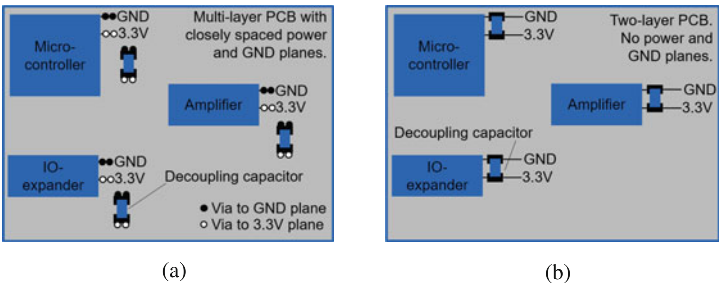

Fig. 16.2 PCB power-supply decoupling. (**a**) PCB with closely spaced power supply and GND planes (global decoupling). (**b**) PCB without closely spaced power supply and GND planes (local decoupling)

primarily be effective for E-fields at any frequency and high-frequency H-fields ($f > 1\,\text{MHz}$), but not for low-frequency H-fields because copper has low relative permeability μ_r' and does not act as a shield against low-frequency magnetic fields.

16.6 Guideline # 5: Route High-Frequency Signals Adjacent to a Plane

Goal = reduced radiated emissions.
High-frequency signals are always to be routed close to an adjacent solid reference plane (GND or power supply plane, which acts in this case as a high-frequency ground). For single-layer designs, use guard traces close to the high-frequency signals where the return currents can flow or GND copper fill.

16.7 Guideline # 6: Control Rise- and Fall-Time

Goal = reduced radiated emissions.
Increase the rise- and fall-time of any digital signal (especially clock signals) as far as possible. A short rise- and fall-time means larger signal bandwidth and, therefore, more high-frequency content in a digital signal, which could lead to radiated emissions or reflections in case of a long transmission line (compared to the wavelength of the high-frequency content).
Rule of thumb: Add a series resistor (typically 33 Ω) close to the driver's output to reduce the rise- and fall-time.
Rule of thumb: The highest significant frequency content f_{max} [Hz] in a digital signal does not depend on the first harmonic (fundamental frequency). Instead, it depends on the rise- and fall-time:

$$f_{max} \approx \frac{0.35}{t_{10\%-90\%}} \tag{16.1}$$

where

$t_{10\%-90\%}$ = rise- and/or fall-time (whichever is smaller) from 10 to 90 % of the slope of a digital signal in [sec]

Rule of thumb: Every PCB trace of length longer than $\lambda/10$ [m] should be considered a transmission line and no longer as a simple interconnection (where λ [m] is the wavelength). This means that such a trace should be laid out with controlled impedance Z_0 [Ω]. In other words, avoid impedance changes or discontinuities along the PCB trace, as these impedance changes or discontinuities could lead to reflections or ringing. Reflections and ringing affect the signal integrity and lead to increased electromagnetic radiation.

16.8 Guideline # 7: Keep Clock lines As Short As Possible

Goal = reduced radiated emissions.
To paraphrase the honorable Henry W. Ott: “Get paranoid about clock routing!” [4]. Take care about clock lines and their return current paths and keep them as short as possible!

16.9 Guideline # 8: Fill Top and Bottom Layers with Circuit GND

Goal = reduced radiated emissions.
Fill the top and bottom layer of a PCB with a solid ground plane around the signals (copper area) and metalize the PCB edges (Fig. 16.3). This helps minimize radiated emission because the filled GND areas at the top and bottom layers shield inner-layer signals and prevent radiation. Moreover, the filled copper areas help maintain a low impedance return current path and short current loops.

Important: do not forget to place a grid of ground stitching vias throughout the whole PCB (otherwise, some small copper islands will start to radiate and you will get more radiation than without the solid copper fill)! This is very important! The rule of thumb in this section below presents a method of determining the maximum distance between stitching vias.

In addition, plated PCB outside edges (connected to circuit GND) help prevent the inner PCB layers from radiating. The plated PCB edges also help increase the cooling efficiency of a PCB because there is an additional copper surface where heat exchange can occur. The additional costs for metalized PCB edges are low.
Rule of thumb: When filling top and bottom layers with ground (copper pour), it is best practice to add a grid of ground stitching vias over the whole PCB. Otherwise, some small GND copper areas would tend to radiate! The distance between these vias within that grid depends on the highest frequency f_{max} [Hz] on the PCB. Given a signal with wavelength λ [m], it is a rule of thumb that a stub or trace of length $l \geq \lambda/10$ starts to become a problem (regarding radiation) and a trace of length $l < \lambda/20$ will not be a problem (in between $\lambda/10$ and $\lambda/20$ is a gray area [4]). Therefore, the distance between the vias should be shorter than $\lambda/10$ of f_{max} [Hz].

Fig. 16.3 Profile view of a multi-layer PCB with metalized edges

Table 16.1 $\lambda/10$ for a given frequency f_{max} [Hz] and PCB material with ε_r'

Smallest rise/fall-time on the PCB [nsec]	Highest frequency f_{max} on the PCB	$\lambda/10$ for ε_r=3,5 (ceramic+PTFE) [mm]	$\lambda/10$ for ε_r=4,5 (FR-4) [mm]	$\lambda/10$ for ε_r=6 (HF-laminate) [mm]	$\lambda/10$ for ε_r=11 (HF-laminate) [mm]
500	0.7 MHz	22908	20203	17496	12922
200	1.75 MHz	9163	8081	6999	5169
100	3.5 MHz	4582	4041	3499	2584
50.0	7 MHz	2291	2020	1750	1292
20.0	17.5 MHz	916	808	700	517
10.0	35 MHz	458	404	350	258
5.00	70 MHz	229	202	175	129
2.00	175 MHz	92	81	70	52
1.00	350 MHz	46	40	35	26
0.50	700 MHz	23	20	17	13
0.20	1.75 GHz	9.2	8.1	7.0	5.2
0.10	3.5 GHz	4.6	4.0	3.5	2.6
0.05	7 GHz	2.3	2.0	1.7	1.3
0.02	17.5 GHz	0.9	0.8	0.7	0.5
0.01	35 GHz	0.5	0.4	0.3	0.3

The wavelength λ [m] of a sinusoidal signal running through a PCB signal trace is according to Eq. 4.2:

$$\lambda = \frac{c}{f \cdot \sqrt{\varepsilon_r'}} \tag{16.2}$$

where

$c = 1/(\sqrt{\mu_0 \varepsilon_0}) = 2.998 \cdot 10^8$ m/s = speed of light
f = frequency of a sinusoidal signal in [Hz]
ε_r' = relative permeability of the PCB material (typical $\varepsilon_r' = 4.5$ for FR4)

However, how to determine f_{max} [Hz] or $\lambda/10$ [m], respectively? Usually, the highest frequency f_{max} [Hz] on a board can be found in the digital signals, e.g., the clock signals:

$$f_{max} = \frac{0.35}{t_{10\%-90\%}} \tag{16.3}$$

where

- $t_{10\%-90\%}$ = rise- and/or fall-time (whichever is smaller) from 10 to 90 % of the slope of a digital signal in [sec]

Table 16.1 shows example values of high-frequency digital signals rise-/fall-time, its corresponding highest frequency content f_{max} [Hz] and $\lambda/10$-values (as mentioned above, the recommended distance between the vias of the grid of vias is $< \lambda/10$).

16.10 Guideline # 9: Add Stitching Vias Around High-Speed Signal Vias

Goal = reduced radiated emissions.
Imagine the following scenario, a high-speed signal changes from one plane to another plane of a PCB. In order to minimize ground bounce, the return current path impedance should be minimized [1]. There are these two options, depending on the return current path:

1. **Identical return current reference nets.** If the two planes have the same reference net with identical electrical potential (e.g., GND), add two or three *stitching vias* (between the reference planes) close to high-speed signal via. These stitching vias help keep current loops and, therefore, the inductance as small as possible.
2. **Different return current reference nets.** If the two reference planes are DC isolated, ensure that the two reference planes are coupled with the lowest impedance possible. This can be achieved with the thinnest possible dielectric layer between them (see Fig. 16.4).

16.11 Guideline # 10: Add a Capacitor Close to Every Pin of a Connector

Goal = reduced radiated emissions, increased immunity.
Filtering of signals directly at the connector is very important! This helps increase ESD immunity to a PCB, reduce radiated emissions, and increase immunity to coupled burst signals on IO-cables. Every signal or power supply line which enters or leaves a PCB needs a filter element, e.g., a ceramic capacitor. One side of the capacitor should be connected close to the connector pin, the other pin tied to the ground plane with low inductance. Table 16.2 proposes some common capacitor values depending on the signal's data rate.
Hint: Signal lines which leave a device (e.g., a connector which people can touch with their hands) is exposed to ESD (±2 kV, ±4 kV, ±6 kV, ±8 kV). In this case,

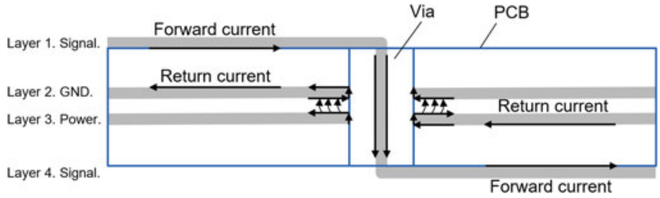

Fig. 16.4 Return current path in case of a reference plane change [1]

Table 16.2 Capacitors of IO-filters vs. data rate (rule of thumb)

Data rate	Rule-of-thumb for EMC, EMI, ESD capacitors as IO-filters
100 kBit/s	< 100 pF
1 MBit/s	< 22 pF
10 MBit/s	< 10 pF
100 MBit/s	TVS diodes with C < 1 pF
>1 GBit/s	TVS diodes with C < 0.1 pF

use capacitors with a high voltage rating (e.g., $>$ 250 V, depending on capacitance and ESD test voltage and other components involved, e.g., like ferrite beads between connector pin and capacitor or conductor length).

16.12 Guideline # 11: Connect Circuit GND to Chassis at IO Area

Goal = reduced radiated emissions, increased immunity.
Bound a circuit GND to chassis at the area where a cable leaves/enters the chassis. Connect it with very low impedance! It is important that GND and chassis have the same potential in the IO area:

- It prevents unintended radiation, as the GND shows a minimum voltage difference to the chassis (earth).
- It makes the IO-signal filters on a PCB most effective and keeps ESD pulses away from the circuits on the board. Why? Because incoming noise and interference from the cable or connector can directly flow back to the source along chassis and earth.

16.13 Guideline # 12: Lay Cables Along Chassis (GND/Earth)

Goal = reduced radiated emissions.
Whenever possible, lay cables constantly along the chassis. This keeps the electromagnetic field generated by the voltages and currents in the cable at a minimum radiation level. The word *constantly* is essential because when laying out cables constantly along a chassis, there is no change in electrical balance (a change in electrical balance leads to common-mode currents) [3].

16.14 Guideline # 13: Don't Use Cable Shield as Signal Conductor for Low-Frequency Signals

Goal = increased immunity.
The cable shield should not be one of the signal conductors for low-frequency signals because of the potential interference with the ground loop current in the shield.
Hint: This rule does not apply for high-frequency signals, where the signal return current and the noise current are separated by the skin effect within the shield (return current of the high-frequency signal flows on the surface of the inner side of the shield and the noise current on the outer surface).

16.15 Guideline # 14: Cable Shield Grounding on Only One End for Low-Frequency Signals

Goal = increased immunity to electric-fields.
It may lead to problems when laying both ends of a cable shield to ground for low-frequency signals because ground loop currents in the shield could interfere with the signals inside the cable.
Rule of thumb: For shielding of low-frequency ($<100\,\mathrm{kHz}$) signals:

- **Shielding against E-fields:** Lay only one end of the shield to ground (with low impedance) to avoid noise current through the shield (e.g., induced by magnetic fields or ground loop currents).
- **Shielding against H-fields:** Lying only one end of the shield to ground does not protect from H-field interference. If protection against low-frequency H-fields is needed, a shield with relative permeability $\mu_r' \gg 1$ is necessary. Twisting does also help to protect against H-field coupling.

If you are in control of only one side of the cable shield (because at the other end of the cable is an unknown device from various manufacturers), then lay the cable shield to ground with low inductance (no pigtails, use a 360° shield clamp) or implement a hybrid ground. A hybrid ground is a compromise for shielding against high-frequency signals, while minimizing low-frequency ground loop currents in the shield (see Sect. 13.7.4 and Fig. 16.5): Lay the cable shield to ground with a resistor (to reduce the ground loop current in the shield for low-frequency signals) and add a parallel capacitor to that resistor (to allow high-frequency signals to flow through the cable shield).

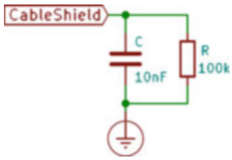

Fig. 16.5 Hybrid cable shield grounding

16.16 Guideline # 15: Cable Shield Grounding on Both Ends for High-Frequency Signals

Goal = increased immunity, reduced magnetic-field emissions.
It is a must to lay both ends of high-frequency signal cable shields to ground with low inductance (no pigtails, use a 360° shield clamp or the like).
Rule of thumb: For high-frequency (>1 MHz) signals:

- The shield can be used as signal return path for high-frequency signals, because the signal return current and the noise current are separated by the skin effect. The induced noise current in the shield helps to cancel out the magnetic field of the outside noise source.
- To reduce the magnetic field emissions from a signal in a shielded cable, the shield has to be laid on ground on both ends [4].

16.17 Guideline # 16: Minimize Loop Area of Signals in Cables

Goal = increased immunity to magnetic-fields.
The best way to protect a signal from magnetic fields is to reduce the current loop area A [m^2] [4]. Minimizing the loop area A [m^2] in case of a cable means twisting the wires of the forward and return current or using neighbor conductors for the forward and return current in flat ribbon cables.

16.18 Guideline # 17: Avoid Electrical Balance Changes

Goal = reduced radiated emissions.
A change from electrically balanced to unbalanced and vice versa (mode conversion) is called an electrical balance change (balanced and unbalanced transmission lines are topic in Sect. 7.9). Electrical balance changes lead to common-mode currents. And we do not want common-mode currents! Common-mode currents are often the cause for unintended radiated emissions. Thus, mode conversions should be avoided. The maximum common-mode voltage $V_{CM,max}$ [V] generated due to

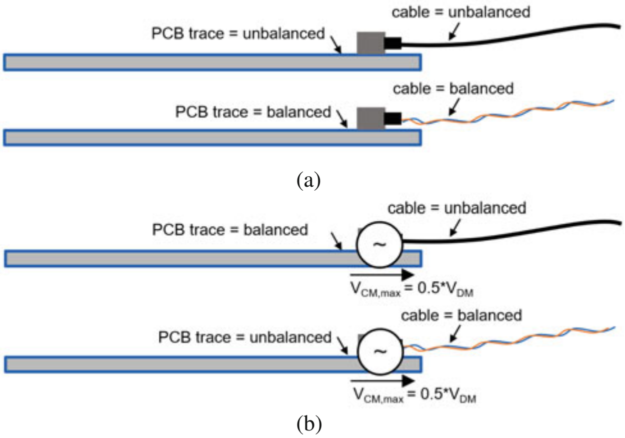

Fig. 16.6 Good and bad examples of PCB-to-cable-interconnections. (**a**) Good examples. (**b**) Bad examples

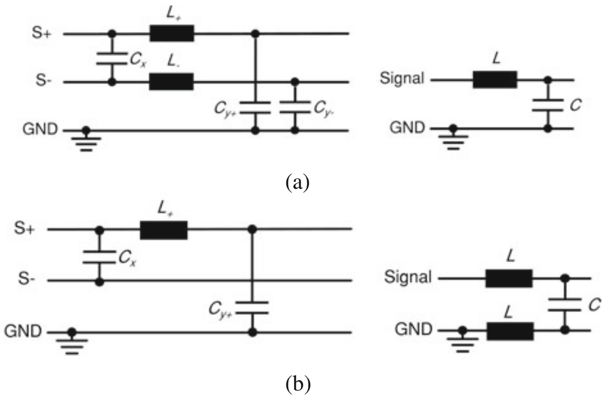

Fig. 16.7 Good and bad examples of how to filter balanced and unbalanced signal lines. (**a**) Good examples. (**b**) Bad examples

mode change is $V_{CM,max} = \Delta h_{max} V_{DM} = 0.5 V_{DM}$ [2], where h is the imbalance factor (Sect. 12.2.1) and Δh_{max} is the maximum imbalance factor change.

Avoiding electrical balance changes for cables connected to a PCB means:

- If the signal is balanced, stay balanced (twisted pair, flat cable). See Fig. 16.6a.
- If the signal is unbalanced, stay unbalanced (coaxial, multi-layer flat cable). See Fig. 16.6a.

Avoiding electrical balance changes for filters means:

- If a signal is balanced, add the identical filters to both signal lines. See Fig. 16.7a.
- If a signal is unbalanced, add only filters to the signal line. See Fig. 16.7a.

Rule of thumb: Send single-ended signals over unbalanced transmission lines and differential signals over balanced transmission lines. Balanced and unbalanced transmission lines are explained in Sect. 7.9 and single-ended and differential signal interfaces in Sect. 7.10.

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Appendix A
Constants

For the sake of brevity, we will always represent this number 2.718281828459... by the letter e.

—Leonhard Euler

This chapter contains useful physical and mathematical *constants* for EMC design engineers.

- **Speed of light.** $c = \frac{1}{\sqrt{\varepsilon_0 \mu_0}} = 2.998 \cdot 10^8$ m/sec $\approx 3 \cdot 10^8$ m/sec.
- **Euler's number.** $e = 2.718281828459$
- **Pi.** $\pi = 3.141592653589793$
- **Permittivity of vacuum.** $\varepsilon_0 = 8.854 \cdot 10^{-12}$ F/m.
- **Permeability of vacuum.** $\mu_0 = 4\pi \cdot 10^{-7}$ H/m $= 12.57 \cdot 10^{-7}$ H/m.
- **Intrinsic impedance of vacuum.** $\eta_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 376.7303\,\Omega \approx 377\,\Omega$.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7

Appendix B
Complex Numbers

The shortest path between two truths in the real domain passes through the complex domain.

—Jacques Hadamard

This chapter does shortly introduce the topic of *complex numbers*, which is commonly used in the field of electrical engineering.

B.1 Definition and Notation

Complex numbers and variables are noted with an underline throughout this book. The complex number $\underline{c}$ has the form:

$$\underline{c} = a + jb \tag{B.1}$$

where

$a = \mathrm{Re}(\underline{c})$ = real number, real part of the complex number $\underline{c}$
$b = \mathrm{Im}(\underline{c})$ = real number, imaginary part of the complex number $\underline{c}$
$j = \sqrt{-1}$

B.2 Cartesian Complex Plane

A complex number $\underline{c}$ can be seen as a point in a Cartesian coordinate system with the coordinates (a,b). The horizontal axis presents the real part and the vertical axis the imaginary part (see Fig. B.1). The Cartesian form is noted as:

$$\underline{c} = a + jb \tag{B.2}$$

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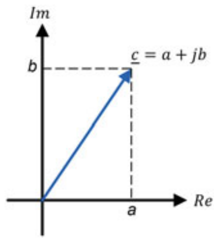

Fig. B.1 Cartesian complex plane

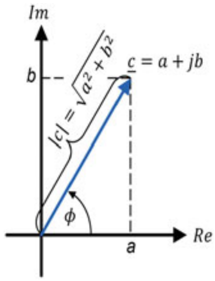

Fig. B.2 Polar complex plane

where

$a = \mathrm{Re}(\underline{c}) =$ real part of $\underline{c} =$ horizontal coordinate
$b = \mathrm{Im}(\underline{c}) =$ imaginary part of $\underline{c} =$ vertical coordinate

B.3 Polar Complex Plane

A complex number $\underline{c}$ can be represented by the polar form with a defined amplitude $|\underline{\mathbf{c}}|$ and angle ϕ [rad]: the polar coordinates $(|\underline{c}|, \phi)$ (see Fig. B.2). The polar form—by using the Euler's formula—is also known as phasor notation and can be written as:

$$\underline{c} = |\underline{c}| \cdot e^{j\phi} \tag{B.3}$$

where

$|\underline{c}| = \sqrt{a^2 + b^2} =$ magnitude or amplitude or absolute value of $\underline{c}$
$\phi = \arctan 2(b, a) =$ angle or phase of $\underline{c}$
$e = 2.71828 =$ Euler's number
$j = \sqrt{-1}$

B.4 Complex Conjugate

The complex conjugate of $\underline{c} = a + jb$ is noted as $\underline{c}^*$ in this book and it is defined as:

$$\underline{c}^* = a - jb \tag{B.4}$$

B.5 Mathematical Operations with Complex Numbers

If we define two complex numbers as $\underline{c}_1 = a_1 + jb_1$ and $\underline{c}_2 = a_2 + jb_2$, we can write:

$$k \cdot \underline{c}_1 = k \cdot (a_1 - jb_1) = ka_1 - jkb_1 \tag{B.5}$$

$$\underline{c}_1 + \underline{c}_2 = a_1 + jb_1 + a_2 + jb_2 = (a_1 + a_2) + j(b_1 + b_2) \tag{B.6}$$

$$\underline{c}_1 - \underline{c}_2 = a_1 + jb_1 - (a_2 + jb_2) = (a_1 - a_2) + j(b_1 - b_2) \tag{B.7}$$

$$\underline{c}_1 \cdot \underline{c}_2 = (a_1 + jb_1) \cdot (a_2 + jb_2) = (a_1 a_2 - b_1 b_2) + j(a_1 b_2 + a_2 b_1) \tag{B.8}$$

B.6 Impedances in the Complex Plane

Figure B.3 presents the impedance of resistors R, inductors L, and capacitors C in the Cartesian complex plane.

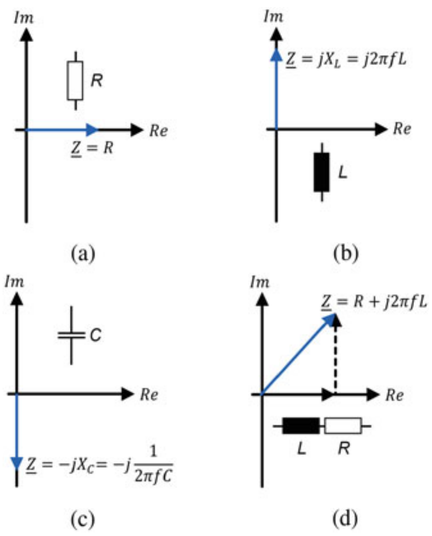

Fig. B.3 *RLC* impedances in the Cartesian complex plane. (**a**) Resistance in the Cartesian complex plane. (**b**) Inductance in the Cartesian complex plane. (**c**) Capacitance in the Cartesian complex plane. (**d**) Resistance and inductance in series

Appendix C
Vectors

The laws of nature are but the mathematical thoughts of God.

—Euclid, father of geometry

This chapter does shortly introduce the topic of the *Euclidean vector geometry*. Euclidean vector geometry is needed for electromagnetism, where we have to deal with vector fields.

C.1 Vector Definition

A vector is a directed line segment with two properties: a direction and a magnitude. Figure C.1 shows a vector $\overrightarrow{A}$ with its projection vectors $\overrightarrow{A_x}$, $\overrightarrow{A_y}$, $\overrightarrow{A_z}$ in a rectangular coordinate system with the unit vectors $\overrightarrow{a_x}$, $\overrightarrow{a_y}$, $\overrightarrow{a_z}$ [1].

C.2 Sum of Vectors

The sum of vectors $\overrightarrow{A}$ and $\overrightarrow{B}$ is shown in Fig. C.2. Vector addition is commutative [1]; this means:

$$\overrightarrow{A} + \overrightarrow{B} = \overrightarrow{B} + \overrightarrow{A} \tag{C.1}$$

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7

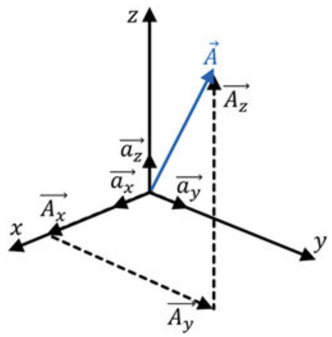

Fig. C.1 A single vector in a rectangular coordinate system

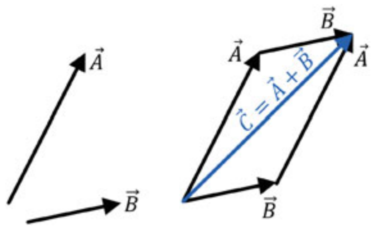

Fig. C.2 Sum of two vectors

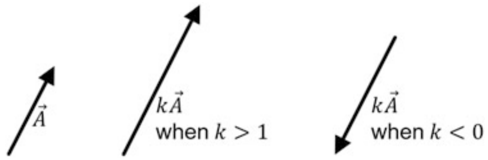

Fig. C.3 Vector multiplication with a scalar

C.3 Vector Multiplication with Scalars

When multiplying vector $\overrightarrow{A}$ with scalar k (a scalar has only a magnitude but no direction), vector $\overrightarrow{A}$ changes its magnitude. Moreover, in the case of a negative scalar, vector $\overrightarrow{A}$ also changes its direction (direction is reversed, see Fig. C.3) [1].

C.4 Vector Dot Product

The *dot product*—also named scalar product—is commutative, and it is defined as [1]:

$$\overrightarrow{A} \cdot \overrightarrow{B} = \overrightarrow{B} \cdot \overrightarrow{A} = |\overrightarrow{A}||\overrightarrow{B}|\cos(\phi_{AB}) \tag{C.2}$$

where

$|\overrightarrow{A}|$ = magnitude of vector $\overrightarrow{A}$

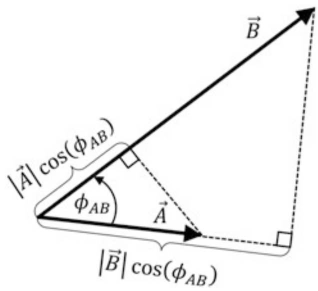

Fig. C.4 The projections of used for the dot product

$|\overrightarrow{B}|$ = magnitude of vector $\overrightarrow{B}$
ϕ_{AB} = smallest angle between vectors $\overrightarrow{A}$ and $\overrightarrow{B}$ in [rad]

The result of the vector dot product is a scalar (Fig. C.4). The scalar product can be interpreted as product of the magnitude of vector $|\overrightarrow{A}|$ and the magnitude of the projection of vector $\overrightarrow{B}$ onto vector $\overrightarrow{A}$ or the scalar product can be interpreted as product of the magnitude of vector $|\overrightarrow{B}|$ and the magnitude of the projection of vector $\overrightarrow{A}$ onto vector $\overrightarrow{B}$.

C.5 Vector Cross Product

The vector cross product of vectors $\overrightarrow{A}$ and $\overrightarrow{B}$ is defined as [1]:

$$\overrightarrow{A} \times \overrightarrow{B} = -\overrightarrow{B} \times \overrightarrow{A} = |\overrightarrow{A}||\overrightarrow{B}|\sin(\phi_{AB})\overrightarrow{a_n} \tag{C.3}$$

where

$|\overrightarrow{A}|$ = magnitude of vector $\overrightarrow{A}$
$|\overrightarrow{B}|$ = magnitude of vector $\overrightarrow{B}$
ϕ_{AB} = smallest angle between vectors $\overrightarrow{A}$ and $\overrightarrow{B}$ in [rad]
$\overrightarrow{a_n}$ = a unit vector which is normal to the plane which contains $\overrightarrow{A}$ and $\overrightarrow{B}$

The direction of the vector cross product $\overrightarrow{A} \times \overrightarrow{B}$ is defined with the right-hand rule (Fig. C.5): point the thumb into the direction of the first vector of the cross product, then point the forefinger into the direction of the second vector of the cross product, and finally, the middle finger points into the direction of the cross product $\overrightarrow{a_n}$. The right-hand rule for $\overrightarrow{Z} = \overrightarrow{X} \times \overrightarrow{Y}$ is shown in Fig. C.6.

Fig. C.5 The cross product of two vectors

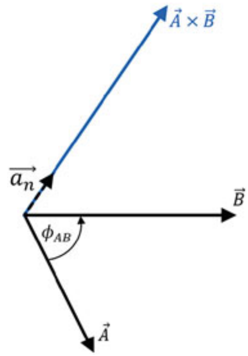

Fig. C.6 The right-hand rule illustrated on the 200 francs Swiss banknote [2]

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Appendix D
Electromagnetism

Faraday is, and must always remain, the father of that enlarged science of electromagnetism.

—James Clerk Maxwell

This chapter contains helpful content and definitions for EMC design engineers on the subject of *electromagnetism*. The most basic electromagnetic definitions and physical quantities are presented.

D.1 Voltage

Voltage is the difference in electric potential between two points in an electric circuit or system. In case of sinusoidal voltage waves, there is often the complex phasor notation used [4]:

$$\underline{V} = |V|e^{j(\omega t+\phi)} = \hat{V}e^{j(\omega t+\phi)} \tag{D.1}$$

where $\omega = 2f\pi$ in [rad/sec], $|V| = \hat{V}$ = magnitude/peak/amplitude of the voltage $\underline{V}$ in [V], t is the time in [sec], and ϕ is an arbitrary angle in [rad].

[V]. The SI[1] unit of an electric potential difference is V Volts [V].
[dBV] $= 20\log_{10}\left(\frac{V}{1\,\mathrm{V}}\right)$. Decibel value of voltage V [V] with reference to 1 V.
[**dBµV**] $= 20\log_{10}\left(\frac{V}{1\,\mu\,\mathrm{V}}\right)$. Decibel value of voltage V [V] with reference to 1 µ V.

[1] SI = The International System of Units is the modern form of the metric system. It is the only system of measurement with an official status in nearly every country in the world.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7

D.2 Current

An electric *current* is a stream of charged particles, such as electrons or ions, moving through an electrical conductor (e.g., wire or PCB trace) or space (e.g., air). In the case of sinusoidal current waves, there is often the complex phasor notation used [4]:

$$\underline{I} = |I|e^{j(\omega t+\phi)} = \hat{I}e^{j(\omega t+\phi)} \tag{D.2}$$

where $\omega = 2f\pi$ in [rad/sec], $|I| = \hat{I}$ = magnitude/peak/amplitude of the current $\underline{I}$ in [A], t is the time in [sec], and ϕ is an arbitrary angle in [rad].

[A]. The SI unit of the electric current I is Ampere [A].
[dBA] $= 20\log_{10}\left(\frac{I}{1\,\mathrm{A}}\right)$. Decibel value of current I [A] with reference to 1 A.
[**dBµA**] $= 20\log_{10}\left(\frac{I}{1\,\mu\,\mathrm{A}}\right)$. Decibel value of current I [A] with reference to 1 µ A.

D.3 Impedance

Electrical *impedance* $\underline{Z}$ [Ω] is the electrical property of a device (e.g., resistor, filter, antenna) which describes how voltage and current interact in that device. The impedance is defined as a complex number with a real part R (resistance) and an imaginary part X (reactance).

$$\underline{Z} = R + jX \tag{D.3}$$

The unit of R and X is Ohm [Ω]. In the case of direct current, the reactance X [Ω] is zero and therefore $\underline{Z} = R$.
Ohm's law states:

$$\underline{Z} = \frac{\underline{V}}{\underline{I}} \tag{D.4}$$

where $\underline{V}$ [V] is the complex voltage across a device with impedance $\underline{Z}$ [Ω] and $\underline{I}$ [A] the complex current through impedance $\underline{Z}$. For direct current (DC), Ohm's law can be simplified to:

$$R = \frac{V}{I} \tag{D.5}$$

[**Ω**]. Impedance $\underline{Z}$ is a complex number with the derived SI unit Ohm [Ω].

D.4 Power

Power P [W] is the amount of energy transferred or converted per unit time. Electric power P [W] is the rate (per unit time) at which an electric circuit transfers electrical energy E [Ws, J]. Electric power is a function of voltage V [V] and current I [A]:

$$P = V_{RMS} \cdot I_{RMS} \tag{D.6}$$

where V_{RMS} is the root mean square voltage in [V] and I_{RMS} the root mean square current in [A]. The active electrical power P [W] that is converted in a component with an impedance of resistance R [Ω] can be calculated like this:

$$P = \frac{V_{RMS}^2}{R} = I_{RMS}^2 \cdot R \tag{D.7}$$

In the phasor domain—with $\underline{V} = \hat{V} e^{j(\omega t + \phi_V)}$ and $\underline{I} = \hat{I} e^{j(\omega t + \phi_I)}$—the average power is defined as [4]:

$$P = \frac{1}{2} \mathrm{Re}(\underline{V} \cdot \underline{I}^*) \tag{D.8}$$

where $\underline{I}^*$ [A] is the complex conjugate of $\underline{I}$ [A] and the factor $\frac{1}{2}$ appears because $\underline{V}$ [V] and $\underline{I}$ [I] represent peak values.

[W]. The SI unit of power is Watt [W] or Joule per second [J/s].
[**dBW**] $= 20 \log_{10}\left(\frac{P}{1\,\mathrm{W}}\right)$. Decibel value of power P [W] with reference to 1 W.
[**dBm**] $= 20 \log_{10}\left(\frac{P}{1\,\mathrm{mW}}\right)$. Decibel value of power P [W] with reference to 1 mW.

D.5 Complex Permittivity

The *permittivity* $\underline{\varepsilon}$ [F/m] is a measure of the electric polarizability of a dielectric. Materials with high permittivity polarize more in response to an applied electric field $\overrightarrow{E}$ [V/m] than materials with low permittivity. Thus, a material with a large dielectric constant (permittivity) is capable of storing more energy. The electric displacement field $\overrightarrow{D}$ [C/m^2] represents the distribution of electric charges in a given medium resulting from the presence of an electric field $\overrightarrow{E}$ [4]:

$$\overrightarrow{D} = \underline{\varepsilon}\,\overrightarrow{E} \tag{D.9}$$

Permittivity $\underline{\varepsilon}$ [F/m] is not a constant, as it can vary with the frequency of the field applied, the field strength of the applied field, humidity, temperature, and other parameters.

The relative permittivity $\underline{\varepsilon}_r$ is defined as the ratio of the absolute permittivity $\underline{\varepsilon}$ [F/m] to the permittivity of vacuum ε_0 [F/m]:

$$\underline{\varepsilon}_r = \frac{\underline{\varepsilon}}{\varepsilon_0} \tag{D.10}$$

The relative permittivity is directly related to the electric susceptibility $\underline{\chi}_e$ by [3]:

$$\underline{\chi}_e = \underline{\varepsilon}_r - 1 \tag{D.11}$$

otherwise written as:

$$\underline{\varepsilon} = \underline{\varepsilon}_r \varepsilon_0 = (1 + \underline{\chi}_e) \cdot \varepsilon_0 \tag{D.12}$$

Permittivity can be written as a complex value [2]:

$$\underline{\varepsilon} = \underbrace{\underline{\varepsilon}_r}_{\substack{Complex\\relative\\permittivity}} \cdot \underbrace{\varepsilon_0}_{\substack{Vacuum\\permittivity}} = (\underbrace{\varepsilon_r'}_{\substack{Real\\part\\energy\\storage}} - j \underbrace{\varepsilon_r''}_{\substack{Imagianry\\part\\dielectric\\loss}}) \cdot \underbrace{\varepsilon_0}_{\substack{Vacuum\\permittivity}} = \varepsilon' - j\varepsilon'' \tag{D.13}$$

where ε' represents the ability to store energy in a medium when an external electric field is applied and ε'' the energy loss—the energy dissipated when a medium is under the influence of an external electric field.

The product of the signal's angular frequency ω [1/rad] and the material's loss factor $\varepsilon'' = \varepsilon_r''\varepsilon_0$ [F/m] is equivalent to the specific conductance of the material σ [S/m] [5]:

$$\sigma = \omega\varepsilon'' \tag{D.14}$$

$$\varepsilon'' = \frac{\sigma}{\omega} \tag{D.15}$$

This means that for an ideal, lossless insulator ($\sigma = 0$): $\varepsilon'' = 0$ and $\underline{\varepsilon} = \varepsilon' = \varepsilon_r\varepsilon_0$.

D.6 Complex Permeability

Permeability $\underline{\mu}$ [H/m] is the measure of magnetization that a material obtains in response to an applied magnetic field. In electromagnetism, the term magnetic field is used for two distinct but closely related vector fields denoted by the symbols $\overrightarrow{H}$ [A/m] (magnetic field) and $\overrightarrow{B}$ [T] (magnetic flux density). However, $\overrightarrow{H}$ [A/m] and $\overrightarrow{B}$ [T] differ in how they account for magnetization and the relation between these two magnetic field vectors is given as [4]:

$$\overrightarrow{B} = \underline{\mu}\,\overrightarrow{H} \tag{D.16}$$

Permeability is not a constant, as it can vary with frequency of the applied magnetic field, the field strength of the applied field, humidity, temperature, and other parameters.

The relative permeability $\underline{\mu}_r$ is the ratio of the absolute permeability $\underline{\mu}$ [H/m] to the permeability of vacuum μ_0 [H/m]:

$$\underline{\mu}_r = \frac{\underline{\mu}}{\mu_0} \tag{D.17}$$

Relative permeability is directly related to magnetic susceptibility $\underline{\chi}_m$ by [3]:

$$\underline{\chi}_m = \underline{\mu}_r - 1 \tag{D.18}$$

otherwise written as:

$$\underline{\mu} = \underline{\mu}_r \mu_0 = (1 + \underline{\chi}_m) \cdot \mu_0 \tag{D.19}$$

Permeability can be written as complex value [2]:

$$\underline{\mu} = \underbrace{\underline{\mu}_r}_{\substack{Complex\\relative\\permeability}} \cdot \underbrace{\mu_0}_{\substack{Vacuum\\permeability}} = (\underbrace{\mu_r'}_{\substack{Real\\part\\energy\\storage}} - j \underbrace{\mu_r''}_{\substack{Imagianry\\part\\magnetic\\loss}}) \cdot \underbrace{\mu_0}_{\substack{Vacuum\\permeability}} = \mu' - j\mu'' \tag{D.20}$$

where μ' [H/m] represents the ability to store energy in a medium when an external magnetic field is applied and μ'' [H/m] the energy loss (the energy dissipated when a medium is under the influence of an external magnetic field).

D.7 Electric Field

An *electric field* is a physical field that surrounds each electric charge and exerts a force on all other charges in the field. The electrostatic force is a vector quantity, and so is the electric field $\overrightarrow{E}$ [V/m]. The electric field $\overrightarrow{E}$ [V/m] at a specific position is equal to the electrostatic force $\overrightarrow{F}$ [N] (Coulomb force) which is a vanishingly small positive test charge q [C] held at that point experiences [4]:

$$\overrightarrow{E} = \frac{\overrightarrow{F}}{q} \tag{D.21}$$

where the unit of $\overrightarrow{E}$ is volt per meter [V/m], $\overrightarrow{F}$ is in Newton [N], and q is in Coulomb [C].

[V/m]. The derived SI unit for an electric field E is Volts per meter [V/m].
[dBV/m] $= 20\log_{10}\left(\frac{E}{1\,\text{V/m}}\right)$. Decibel value of electric field strength E [V/m] with reference to 1 V/m.
[dBμV/m] $= 20\log_{10}\left(\frac{E}{1\,\mu\,\text{V/m}}\right)$. Decibel value of electric field strength E [V/m] with reference to 1 μ V/m.

D.8 Electric Flux

The *electric flux* Φ_E [Vm] through a surface is the surface integral of the normal component of the electric field $\overrightarrow{E}$ [V/m] over that surface $\overrightarrow{A}$ [m^2]. Electric flux is proportional to the total number of electric field lines going through a surface. If the electric field is uniform, the electric flux through a surface of vector area $\overrightarrow{A}$ [m^2] is [4]:

$$\Phi_E = \overrightarrow{E} \cdot \overrightarrow{A} = |E| \cdot |A| \cdot \cos\theta \tag{D.22}$$

where $|E|$ [V/m] is the magnitude of the electric field (the electric field vector $\overrightarrow{E}$), $|A|$ [m^2] is the area of the surface, and θ [rad] is the angle between the electric field vector and the normal (perpendicular) of $\overrightarrow{A}$ [m^2].

[Vm]. The derived SI unit of electric flux Φ_E is the [Vm].

D.9 Electric Flux Density

In a dielectric material, the presence of an electric field $\overrightarrow{E}$ [V/m] causes the bound charges in the material (atomic nuclei and their electrons) to slightly separate, inducing a local electric dipole moment. The *electric flux density* $\overrightarrow{D}$ [C/m^2]—also called *electric displacement field*—is defined as [4]:

$$\overrightarrow{D} = \varepsilon_0 \overrightarrow{E} + \overrightarrow{P} = \varepsilon_0(1 + \chi_e)\overrightarrow{E} = \varepsilon_0\varepsilon_r' \overrightarrow{E} \tag{D.23}$$

where ε_0 [F/m] is the vacuum permittivity (permittivity of free space), $\overrightarrow{E}$ [V/m] is the electric field intensity, $\overrightarrow{P}$ [C/m^2] is the polarization density vector field (or electric polarization, or simply polarization) that expresses the density of permanent or induced electric dipole moments in a dielectric material, χ_e is the electric susceptibility of the material (a dimensionless proportionality constant that indicates the degree of polarization of a dielectric material in response to an applied electric field), and ε_r' is the complex relative permittivity of the material (dielectric constant). For vacuum: $\chi_e = 0$ and therefore $\varepsilon_r' = 1$.

[**C/m²**]. The derived SI unit for an electric displacement field D is coulombs per square meter [C/m²].

D.10 Displacement Current Density

The *displacement current density* $\overrightarrow{J}_D = \partial \overrightarrow{D}/\partial t$ [A/m²] is defined as the rate of change of the electric flux density $\overrightarrow{D}$ [C/m²]. In other words, the displacement current is the partial derivative in time of the electric flux density $\overrightarrow{J}_D = \partial \overrightarrow{D}/\partial t$ [A/m²]. The displacement current density $\overrightarrow{J}_D$ [A/m²] has the same unit like the electric current density $\overrightarrow{J}$ [A/m²], and it is a source of the magnetic field $\overrightarrow{H}$ [A/m] just as the electric current $\underline{I}$ [A] is. However, the *displacement current density* is not an electric current $\underline{I}$ [A] of moving charges but a time-varying electric field $\partial \overrightarrow{E}/\partial t$. The displacement current density is defined as [4]:

$$\overrightarrow{J}_D = \frac{\partial \overrightarrow{D}}{\partial t} = \underbrace{\varepsilon_0 \frac{\partial \overrightarrow{E}}{\partial t}}_{\substack{Displacement\\ current}} + \underbrace{\frac{\partial \overrightarrow{P}}{\partial t}}_{\substack{Polarization\\ current}} \tag{D.24}$$

where ε_0 [F/m] is the vacuum permittivity (permittivity of free space), $\overrightarrow{E}$ [V/m] is the electric field vector, and $\overrightarrow{P}$ [C/m²] is the polarization vector of the medium.

The term $(\varepsilon_0 \partial \overrightarrow{E}/\partial t)$ refers to the displacement current density $\overrightarrow{J}_D$ [A/m²] in a vacuum, and it does not necessarily come from any actual movement of charges. However, it has an associated magnetic field $\overrightarrow{H}$ [A/m], just as a current $\underline{I}$ [A] due to charge motion. Some authors apply the name *displacement current* to the term $(\varepsilon_0 \partial \overrightarrow{E}/\partial t)$ by itself.

The term $(\partial \overrightarrow{P}/\partial t)$ comes from the change in polarization of the individual molecules of the medium (e.g., a dielectric material). Polarization results when, under the influence of an applied electric field $\overrightarrow{E}$ [V/m], the charges in molecules have moved from a position of exact cancellation. The positive and negative charges in molecules separate, causing an increase in the state of polarization $\overrightarrow{P}$ [C/m²]. A changing state of polarization corresponds to charge movement and so is equivalent to a current, hence the term *polarization current*.

[**A/m²**]. The derived SI unit for the displacement current density J_D is ampere per square meter [A/m²].

D.11 Polarization Density

The polarization density $\vec{P}$ [C/m^2]—also called electric polarization or simply polarization—is the vector field that expresses the density of permanent or induced electric dipole moments in a dielectric material. An external electric field $\vec{E}$ [V/m] that is applied to a dielectric material causes a displacement of bound charged elements. These are elements which are bound to molecules and are not free to move around the material. Positive charged elements are displaced in the direction of the field, and negative charged elements are displaced opposite to the direction of the field. The molecules may remain neutral in charge, yet an electric dipole moment forms.

In general, the dipole moment $\vec{p}$ [Cm] changes from point to point within the dielectric. Hence, the polarization density $\vec{P}$ [C/m^2] of a dielectric inside an infinitesimal volume dV [m^3] with an infinitesimal dipole moment $d\vec{p}$ [Cm] is:

$$\vec{P} = \frac{d\vec{p}}{dV} \tag{D.25}$$

[**C/m^2**]. The derived SI unit for the polarization density P is coulombs per square meter [C/m^2].

D.12 Electric Current Density

The *current density* vector $\vec{J}$ [A/m^2] is defined as a vector whose magnitude is the electric current $\underline{I}$ [A] per cross-sectional area A [m^2] at a given point in space, its direction being that of the motion of the positive charges. The electric current $\underline{I}$ [A] can be defined by the integral of the electrical current density [4]:

$$\underline{I} = \iint_A \vec{J} \cdot d\vec{a} \tag{D.26}$$

where $\iint_A$ is the surface integral over the surface area A [m^2], $\vec{J}$ [A/m^2] is the current density vector which points through that cross-sectional A [m^2], and $d\vec{a}$ [m^2] is the differential cross-sectional area vector.

The current density $\vec{J}$ [A/m^2] in materials with finite electrical resistance is directly proportional to the electric field $\vec{E}$ [V/m] in that medium. The proportionality constant is the specific conductance σ [S/m, 1/(Ωm)] or resistivity ρ [Ωm] of the material [4]:

$$\vec{J} = \sigma\vec{E} = \frac{\vec{E}}{\rho} \tag{D.27}$$

[**A/m²**]. The derived SI unit for the electric current density J is ampere per square meter [A/m²].

D.13 Magnetic Field

A *magnetic field* $\overrightarrow{H}$ [A/m] is a vector field that describes the magnetic influence on moving electric charges, electric currents, and magnetic materials. The magnetic $\overrightarrow{H}$ [A/m] field is defined [4]:

$$\overrightarrow{H} = \frac{1}{\mu_0} \cdot \overrightarrow{B} - \overrightarrow{M} \tag{D.28}$$

where $\mu_0 = 12.57 \cdot 10^{-7}$ H/m is the vacuum permeability, $\overrightarrow{B}$ [T] is the magnetic flux density vector field, and $\overrightarrow{M}$ [A/m] is the magnetization vector (how strongly a region of material is magnetized). In a vacuum, the $\overrightarrow{B}$ [T] and $\overrightarrow{H}$ [A/m] fields are related through the vacuum permeability: $\overrightarrow{B}/\mu_0 = \overrightarrow{H}$; but in a magnetized material, the terms differ by the material's magnetization $\overrightarrow{M}$ [A/m] at each point. The relation of the magnetic field $\overrightarrow{H}$ [A/m] and the magnetic flux density $\overrightarrow{B}$ [T] is often described with the following formula:

$$\overrightarrow{H} = \frac{1}{\mu_0 \cdot \underline{\mu}_r} \cdot \overrightarrow{B} \tag{D.29}$$

where $\underline{\mu}_r$ is the complex relative permeability of the material which is penetrated by the magnetic flux Φ_M [Wb].

[A/m]. The derived SI unit for a magnetic field H is Ampere per meter [A/m].
[dBA/m] $= 20\log_{10}\left(\frac{H}{1\,\text{A/m}}\right)$. Decibel value of magnetic field strength H [A/m] with reference to 1 A/m.
[dBμA/m] $= 20\log_{10}\left(\frac{H}{1\,\mu\,\text{A/m}}\right)$. Decibel value of magnetic field strength H [A/m] with reference to 1 μ A/m.

D.14 Magnetic Flux

The *magnetic flux* Φ_M [Wb] through a surface is the surface integral of the normal component of the magnetic field $\overrightarrow{B}$ [T] over that surface. It is analogous to electric current $\underline{I}$ [A] in an electric circuit. If the magnetic field is uniform, the magnetic flux passing through a surface of vector area $\overrightarrow{A}$ [m²] is [4]:

$$\Phi_M = \overrightarrow{B} \cdot \overrightarrow{A} = |B| \cdot |A| \cdot \cos\theta \tag{D.30}$$

where $|B|$ [T] is the magnitude of the magnetic field (the magnetic flux density vector $\overrightarrow{B}$), $|A|$ [m^2] is the area of the surface, and θ [rad] is the angle between the magnetic field vector and the normal (perpendicular) of $\overrightarrow{A}$ [m^2].

[Wb]. The derived SI unit of magnetic flux Φ_M is the Weber [Wb].

D.15 Magnetic Flux Density

The number of magnetic flux lines Φ_M [Wb] that go through the unit area perpendicular to the magnetic field $\overrightarrow{H}$ [A/m] is called *magnetic flux density* $\overrightarrow{B}$ [T]. The relation of the magnetic flux density $\overrightarrow{B}$ [T] and the magnetic field $\overrightarrow{H}$ [A/m] is often described with the following formula [4]:

$$\overrightarrow{B} = \mu_0 \left(\overrightarrow{H} + \overrightarrow{M}\right) \tag{D.31}$$

where μ_0 [H/m] is the vacuum permeability and $\overrightarrow{M}$ [A/m] is the magnetization vector. In a vacuum, $\overrightarrow{B}$ [T] and $\overrightarrow{H}$ [A/m] are proportional, with the multiplicative constant depending on the physical units. Inside a material, they are different and not proportional to each other (non-linear dependency). A different notation for the magnetic flux density vector $\overrightarrow{B}$ [B] is:

$$\overrightarrow{B} = \mu_0 \underline{\mu}_r \overrightarrow{H} \tag{D.32}$$

where $\underline{\mu}_r$ is the complex relative permeability of the material which is penetrated by the magnetic flux.

[T]. Tesla is a derived SI unit of the *magnetic induction* (magnetic flux density). One Tesla is equal to one Weber per square meter [Wb/m^2]. According to the Lorentz force law (see Appendix E.5), a particle carrying a charge of one Coulomb [C] that moves perpendicularly through a magnetic field of one Tesla [T] at a speed of one meter per second [m/sec] experiences a force with magnitude of one Newton [N].

$$\mathrm{T} = \frac{\mathrm{V{\cdot}s}}{\mathrm{m}^2} = \frac{N}{\mathrm{A{\cdot}m}} = \frac{\mathrm{J}}{\mathrm{A{\cdot}m}^2} = \frac{\mathrm{H{\cdot}A}}{\mathrm{m}^2} = \frac{\mathrm{Wb}}{\mathrm{m}^2}$$

where T = Tesla, V = Volt, s = second, m = meter, N = Newton, A = Ampere, J = Joule, H = Henry, and Wb = Weber. 1 T = 10 000 G.

[mT] $= 1 \cdot 10^{-3}$T = 1 mT = 10 G.
[pT] $= 1 \cdot 10^{-12}$T = 1 pT $= 1 \cdot 10^{-8}$ G.

[dBpT] $= 20\log_{10}\left(\frac{B}{1\,\mathrm{pT}}\right)$. Decibel value of the magnetic flux density B [T] with reference to 1 pT.

[G]. The Gauss [G] is the unit of measurement of magnetic induction, also known as magnetic flux density. $1\,\mathrm{G} = 1\cdot 10^{-4}\mathrm{T} = 0.1\,\mathrm{mT}$

D.16 Electromagnetic Field

The *electromagnetic field* can be regarded as the combination of an electric field $\vec{E}$ [V/m] and a magnetic field $\vec{H}$ [A/m]. The electric field $\vec{E}$ [V/m] is caused by static charges, whereas the magnetic field $\vec{H}$ [A/m] is caused by moving charges (currents). How charges and currents interact with the electromagnetic field is described by Maxwell's equations and the Lorentz force law (see Appendix E.5). The *Poynting vector*[2] $\vec{S}$ [W/m^2] represents the direction of propagation of an electromagnetic wave. $\vec{S}$ [W/m^2] is defined as the cross product of the two vector fields $\vec{E}$ [V/m] and $\vec{H}^*$ [A/m] [1]:

$$\vec{S} = \vec{E} \times \vec{H}^* \tag{D.33}$$

where $\vec{H}^*$ [A/m] is the complex conjugate of $\vec{H}$ [A/m]. For applications of time-varying fields, it is often desirable to find the *average power density* $\vec{S}_{av}$ [W/m^2]. The time average Poynting vector $\vec{S}_{av}$ [W/m^2] can be written as [1]:

$$\vec{S}_{av} = \frac{1}{2}\,\mathrm{Re}\left(\vec{E} \times \vec{H}^*\right) \tag{D.34}$$

where the factor $\frac{1}{2}$ appears because the $\vec{E}$ [V/m] and $\vec{H}$ [A/m] fields represent peak values.

For a uniform plane wave (electromagnetic field in the far-field: wave impedance $\underline{Z}_w$ [Ω] is equal intrinsic impedance $\underline{\eta}$ [Ω]) in a lossless medium ($\underline{\varepsilon}_r = 1$ and $\underline{\mu}_r = 1$) the calculation of the average power density S_{av} [W/m^2] can be simplified to:

$$S_{av} = \frac{1}{2}|\vec{E}|\cdot|\vec{H}| = \frac{|\vec{E}|^2}{2|\underline{\eta}|} = \frac{|\vec{H}|^2|\underline{\eta}|}{2} \tag{D.35}$$

[2] The Poynting vector represents the directional energy flux (the energy transfer per unit area per unit time) of an electromagnetic field. The SI unit of the Poynting vector is the watt per square meter [W/m^2]. It is named after its discoverer John Henry Poynting who first derived it in 1884.

where $|\overrightarrow{E}|$ [V/m] is the amplitude (magnitude) of the electric field, $|\overrightarrow{H}|$ [A/m] is the amplitude of the magnetic field, and $|\underline{\eta}|$ [Ω] is the intrinsic impedance of where the wave is traveling through. The factor $\frac{1}{2}$ can be omitted for RMS values of $\overrightarrow{E}$ [V/m] and $\overrightarrow{H}$ [V/m]. For free-space $|\underline{\eta}|$ can be set to $\eta_0 = 377\,\Omega$.

[W/m²]. The derived SI unit of the power density S of a radiated electromagnetic field is Watts per meter [W/m²].

[dBmW/m²] $= 20\log_{10}\left(\frac{S}{1\,\mathrm{mW/m^2}}\right)$. Decibel value of the power density S [W/m²] with respect to $1\,\mathrm{mW/m^2}$.

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Appendix E
Maxwell's Equations and Lorentz's Force Law

The special theory of relativity owes its origins to Maxwell's equations of the electromagnetic field.

—Albert Einstein

This chapter presents Maxwell's equations and Lorentz's force law. These formulas are the foundation of classical electromagnetism.

E.1 Gauss' Law for Electricity

Electric field lines diverge from a positive charge and converge on a negative charge; the charge is the source of the electric field. The electric flux Φ_E [Vm] out of any closed surface is proportional to the total charge q [C] enclosed within the surface A [m^2] [1]:

$$\underbrace{\oint_A \vec{E} \cdot d\vec{a}}_{\substack{Electric \\ flux}} = \frac{q}{\varepsilon_0} \tag{E.1}$$

where

$\oint_A$ = closed surface integral over the closed surface area A (e.g., a sphere)
$\vec{E}$ = electric field vector in [V/m]
$d\vec{a}$ = infinitesimal small surface area vector in [m^2]
q = electric charge within the closed surface A in [C]
A = closed surface area around the charge q in [m^2]
$\varepsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
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E.2 Gauss' Law for Magnetism

Magnetic field lines do not diverge, and the field of the magnetic flux density is source-free; there are no magnetic monopoles. The net magnetic flux Φ_B [Wb] out of any closed surface A [m^2] is zero [1]:

$$\underbrace{\oint_A \overrightarrow{B} \cdot d\overrightarrow{a}}_{\substack{Magnetic\\ flux}} = 0 \tag{E.2}$$

where

$\oint_A$ = closed surface integral over the closed surface area A (e.g., a sphere)
$\overrightarrow{B}$ = magnetic flux density vector in [T]
$d\overrightarrow{a}$ = infinitesimal small surface area vector in [m^2]
A = closed surface area in [m^2]

E.3 Faraday's Law

The line integral of the electric field $\overrightarrow{E}$ [V/m] around a closed loop L [m] is equal to the negative of the rate of change of the magnetic flux Φ_B [Wb] through the area enclosed by the loop. This line integral is equal to the generated voltage or electromotive force emf [V] in the loop. Faraday's law is the basis for electric generators [1]:

$$\oint_L \overrightarrow{E} \cdot d\overrightarrow{l} = \underbrace{-\frac{d}{dt} \underbrace{\iint_A \overrightarrow{B} \cdot d\overrightarrow{a}}_{\substack{Magnetic\\ flux}}}_{\substack{Magnetic\\ flux\\ change}} = \text{emf} \tag{E.3}$$

where

$\oint_L$ = closed line integral along a the closed loop L (e.g., a circle)
$\overrightarrow{E}$ = electric field vector in [V/m]
$d\overrightarrow{l}$ = infinitesimal small vector along the closed loop in [m]
$\iint_A$ = surface integral over the surface area A (which is enclosed by L)
$\overrightarrow{B}$ = magnetic flux density vector in [T]
$d\overrightarrow{a}$ = infinitesimal small surface area vector in [m^2]
L = closed loop distance with length in [m]

A = surface area (which is enclosed by L) in [m^2]
emf = electromotive force: the open circuit output voltage developed between two terminals of an electric generator/battery (source) in [V]

E.4 Ampére's Law

In the case of static electric field $\vec{E}$ [V/m] (meaning $\partial \vec{E}/\partial t = 0$), the line integral of the magnetic flux density $\vec{B}$ [T] around a closed loop L [m]—which encloses an area A [m^2]—is proportional to the conduction electric current $\underline{I}_C$ [A] flowing through the loop area [1]:

$$\oint_L \vec{B} \cdot d\vec{l} = \mu_0 \iint_A \left(\vec{J_C} + \vec{J_D}\right) \cdot d\vec{a} \tag{E.4}$$

$$= \iint_A \left(\mu_0 \vec{J}_C + \mu_0 \frac{\partial \vec{D}}{\partial t}\right) \cdot d\vec{a} \tag{E.5}$$

$$= \iint_A \left(\mu_0 \vec{J}_C + \mu_0 \varepsilon_0 \frac{\partial \vec{E}}{\partial t}\right) \cdot d\vec{a} \tag{E.6}$$

$$= \iint_A \mu_0 \vec{J}_C \cdot d\vec{a} + \mu_0 \varepsilon_0 \cdot \underbrace{\frac{d}{dt} \underbrace{\iint_A \vec{E} \cdot d\vec{a}}_{\substack{Electric\\flux}}}_{\substack{Electric\\flux\\change}} \tag{E.7}$$

$$= \mu_0 \underline{I}_C + \mu_0 \varepsilon_0 \frac{d\Phi_E}{dt} \tag{E.8}$$

$$= \mu_0 \left(\underline{I}_C + \underline{I}_D\right) \tag{E.9}$$

where

$\oint_L$ = closed line integral along a the closed loop L (e.g., a circle)
$\vec{B}$ = magnetic flux density vector in [T]
$d\vec{l}$ = infinitesimal small vector along the closed loop in [m]
$\iint_A$ = surface integral over the surface area A (which is enclosed by L)
$\vec{J}_C$ = conduction current density vector in [A/m^2]
$\vec{J}_D$ = displacement current density vector in [A/m^2]
$\mu_0 = 12.57 \cdot 10^{-7}$ H/m = permeability of vacuum
$\varepsilon_0 = 8.854 \cdot 10^{-12}$ F/m = permittivity of vacuum
$\partial \vec{E}/\partial t$ = change of electric field per time in [(V/m)/sec]

$d\overrightarrow{a}$ = infinitesimal small surface area vector in [m^2]
Φ_E = electric flux through the area A in [Vm]
$\underline{I}_C$ = conduction current due to the flow of charges through the area A enclosed by loop line L in [A]
$\underline{I}_D$ = displacement current due to the rate of change of the electric field $\overrightarrow{E}$ through the area A enclosed by loop line L in [A]

E.5 Lorentz's Force Law

The Lorentz force (or electromagnetic force) is the combination of electric and magnetic force on a point charge q [C] due to electromagnetic fields $\overrightarrow{E}$ [V/m] and $\overrightarrow{B}$ [T]. A particle of charge q [C] moving with a velocity $\overrightarrow{v}$ [m/sec] in an electric field $\overrightarrow{E}$ [V/m] and a magnetic field with the magnetic flux density $\overrightarrow{B}$ [T] experiences a force of [1]:

$$\overrightarrow{F} = \underbrace{q\overrightarrow{E}}_{\substack{Electric\\ force}} + \underbrace{q\overrightarrow{v} \times \overrightarrow{B}}_{\substack{Magnetic\\ force}} \tag{E.10}$$

where
$\overrightarrow{F}$ = force vector which is experienced by charge q in [N]
q = electric charge of the (moving) particle in [C]
$\overrightarrow{E}$ = electric field vector in [V/m]
$\overrightarrow{B}$ = magnetic flux density vector in [T]
$\overrightarrow{v}$ = velocity vector in [m/sec]

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Appendix F
Integral Transforms

Fourier is a mathematical poem.

—Lord Kelvin

This chapter presents the most popular integral transforms in a compact format. Integral transforms are used to transform time-domain signals to the frequency domain or control systems and filter design.

F.1 Fourier Series

The *Fourier series* is applied to continuous, periodic signals. The frequency spectrum is discrete. Formulas for the calculation of the complex Fourier coefficients c_n of a signal $x(t)$ are presented below [2]:

$$x(t) = \sum_{n=-\infty}^{\infty} \left(c_n \cdot e^{j\omega_0 nt} \right) \tag{F.1}$$

$$c_n = \frac{1}{T} \cdot \int_0^T \left(x(t) \cdot e^{-j\omega_0 nt} \cdot dt \right) \tag{F.2}$$

$$\omega_0 = 2\pi f_0,\ f_0 = \tfrac{1}{T}$$

where T [sec] is the period length of the signal $x(t)$, f_0 [Hz] is the first harmonic of the signal $x(t)$, ω_0 [rad] is the first harmonic of the signal $x(t)$, $j = \sqrt{-1}$, and n stands for the n-th harmonic frequency [0, 1, 2, ...].

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F.2 Fourier Transform

The *Fourier transform* is applied to continuous, non-periodic signals. The frequency spectrum is continuous. Formulas of the Fourier transform of a signal $x(t)$ and the inverse-Fourier transform are given below [2]:

$$X(\omega) = \int_{-\infty}^{\infty} \left(x(t) \cdot e^{-j\omega t} \cdot dt \right) \tag{F.3}$$

$$x(t) = \frac{1}{2\pi} \cdot \int_{-\infty}^{\infty} \left(X(\omega) \cdot e^{j\omega t} \cdot d\omega \right) \tag{F.4}$$

where $X(\omega)$ is the Fourier transform (spectrum) of $x(t)$, ω is the frequency in [rad], $j = \sqrt{-1}$, and t the time in [sec].

F.3 Discrete Fourier Transform (DFT)

The *discrete Fourier transform* is applied to discrete, periodic signals. The frequency spectrum is discrete. The DFT is also applied to non-periodic signals in the time domain by periodically continuing the non-periodic signals to make them computable with the DFT. The DFT is by far the most common method of modern Fourier analysis. The *Fourier Transform!Fast Fourier Transformation* (FFT) is a fast algorithm for calculating the DFT (if the block length N is a power of two) [1].

$$X[k] = \sum_{n=0}^{N-1} \left(x[n] \cdot e^{-j\frac{2\pi f_s}{N}kn} \right), k = 0, 1, 2, ..., N-1 \tag{F.5}$$

$$x[n] = \frac{1}{N} \cdot \sum_{k=0}^{N-1} \left(X[k] \cdot e^{j\frac{2\pi f_s}{N}kn} \right), n = 0, 1, 2, ..., N-1 \tag{F.6}$$

$$\omega_k = \tfrac{2\pi f_s}{N}k, \; f_s = \tfrac{1}{T_s}$$

where X is the spectrum of x and $X[k]$ is the k-th sample of the spectrum at ω_k. $x[n]$ is the amplitude of the n-th sample of the discrete-time signal, which is sampled with the sampling time T_s [sec] and the sampling frequency f_s [Hz]. In the digital signal processing literature, f_s [Hz] is often set to one and the equations for the DFT and the IDFT are simplified like this:

$$X[k] = \sum_{n=0}^{N-1} \left(x[n] \cdot e^{-j\frac{2\pi}{N}kn} \right), k = 0, 1, 2, ..., N-1 \tag{F.7}$$

$$x[n] = \frac{1}{N} \cdot \sum_{k=0}^{N-1} \left(X[k] \cdot e^{j \frac{2\pi}{N} kn} \right), n = 0, 1, 2, ..., N - 1 \tag{F.8}$$

$$\omega_k = \tfrac{2\pi}{N} k,\ f_s = 1,\ T_s = 1$$

F.4 Discrete-Time Fourier Transform (DTFT)

The *discrete-time Fourier transform* is applied to discrete, non-periodic signals. The frequency spectrum is continuous. The DTFT can be viewed as the form of the DFT when its length N approaches infinity [1].

$$X(\widetilde{\omega}) = \sum_{n=-\infty}^{\infty} \left(x[n] \cdot e^{-j\widetilde{\omega}n} \right) \tag{F.9}$$

$$x[n] = \frac{1}{2\pi} \int_{-\pi}^{\pi} \left(X(\widetilde{\omega}) \cdot e^{j\widetilde{\omega}n} \cdot d\widetilde{\omega} \right) \tag{F.10}$$

$$\widetilde{\omega} = \omega T_s,\ f_s = \tfrac{1}{T_s}$$

were X is the spectrum of x and $x[n]$ is the signal amplitude of the n-th sample of the discrete-time signal, which is sampled with the sampling time T_s [sec] and the sampling frequency f_s [Hz]. $\widetilde{\omega}$ denotes the continuous normalized radian frequency variable $[-\pi ... + \pi]$. $\widetilde{\omega}$ is the product of the radian frequency ω [rad] and the sampling time T_s [sec].

F.5 Laplace Transform

The *Laplace transformation* belongs—like the Fourier analyses—to the group of integral transformations. It is mentioned here for the sake of completeness. The Laplace transform is used for system analysis (e.g., control systems, filters), whereas the Fourier transform is used for signal analysis [1].

$$X(s) = \mathscr{L}\{x(t)\} = \int_{-\infty}^{\infty} \left(x(t) \cdot e^{-st} \cdot dt \right) \tag{F.11}$$

$$s = \sigma + j\omega$$

where X is the Laplace transform of x, t [sec] the time (real variable) and $s = \sigma + j\omega$ is a complex variable with a real part σ and an imaginary part with the radian frequency ω [rad]. In case of $\sigma = 0$, the Laplace transform reduces to the Fourier transform.

F.6 Z-Transform

The unilateral *Z-transform* starts at $n = 0$ (and not at $n = -\infty$, since signals are typically defined to begin at time $n = 0$, and since filters are often assumed to be causal) and is the time-discrete counterpart to the Laplace transform with $z = e^{sT_s} = e^{(\sigma + j\omega)T_s}$ [1].

$$X[z] = \sum_{n=0}^{\infty} \left(x[n] \cdot z^{-n}\right) \tag{F.12}$$

$$z = e^{sT_s}, s = \sigma + j\omega, f_s = \frac{1}{T_s}$$

where X is the Z-transform of x, z is a complex variable, and $x[n]$ is the n-th sample of the discrete-time signal, which is sampled with the sampling time T_s [sec] and the sampling frequency f_s [Hz].

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Appendix G
Material Properties

You cannot be sure that you are right unless you understand the arguments against your views better than your opponents do.

—Milton Friedman

This chapter contains comprehensive data on relevant materials for EMC design engineers. This comprises metals with their conductivity and relative permeability (important properties for shielding applications) and plastics with their relative permittivity. In addition, the relevant components of a PCB are presented and explained.

G.1 Conductor Properties

Metals and their properties play an important role in EMC. For example, it is necessary to know the *specific conductance* σ [S/m] or *resistivity* ρ [Ω/m] to calculate the skin depth. The values in Table G.1 applies for DC (0 Hz) and room temperature (25 °C).

In addition to the electrical conductivity σ [S/m], the relative *permeability* μ_r' of materials is another important property, e.g., when it comes to shielding of low-frequency ($f < 100$ kHz) magnetic fields. The relative magnetic permeability μ_r' of a material tells us how much better this material is able to "conduct" the magnetic flux, or in other words, how big the flux-density B [T] in a material is compared to vacuum (where $\mu_r' = 1$) for a given field strength H [A/m].

Our primary interest lies in *ferromagnetic materials* (soft magnetic) because they can be used for shielding low-frequency magnetic fields. Let us have a look at how magnetic materials are classified:

- **Anti-ferromagnetic.** μ_r' is slightly bigger than 1. The only pure metal which is anti-ferromagnetic is chromium (Cr).
- **Diamagnetic.** μ_r' is slightly smaller than one. Diamagnetic materials are weakly repelled by a magnet. They cannot be magnetized.

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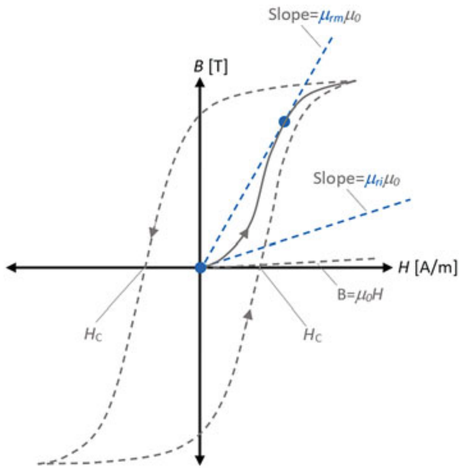

Fig. G.1 Magnetization curve of a ferromagnetic metal. The initial relative magnetic permeability μ'_{ri} describes the relative permeability for low flux densities B [T]. The maximum relative magnetic permeability μ'_{rm} is usually by factor 2...5 (or even more) higher than μ'_{ri}. μ'_{rm} is valid at a single point in the H-B-hysteresis diagram (at this point the change of the magnetic field H [A/m] results in the biggest change in the flux density B [T])

- **Paramagnetic.** μ'_r is slightly bigger than one. Paramagnetic materials are weakly attracted by a magnet. They cannot be magnetized.
- **Ferrimagnetic.** μ'_r is bigger than 1, but much smaller compared to the μ'_r of ferromagnets. Ferrimagnetic materials can be weakly magnetized ("weak magnets").
- **Ferromagnetic.** μ'_r is much bigger than 1. Ferromagnetic materials can be magnetized and used as shielding materials (even against low-frequency magnetic fields) or for building permanent magnets. Ferromagnetic materials can be categorized by their coercivity H_c [A/m] (Fig. G.1). A high coercivity H_c of a material means that the external magnetic field must be high to change the polarization of the magnet.
  - **Soft Magnetic.** Coercivity H_c [A/m] is low (typically < 1 kA/m). Example applications: shielding, transformers, and ferrite cores.
  - **Hard Magnetic.** Coercivity H_c [A/m] is high (typically > 10 kA/m). Example application: permanent magnets.

The data for μ'_r in Table G.1 applies to room temperature (25 °C) and DC (0 Hz). Please note that μ'_r is current I [A], temperature T [°C], and frequency f [Hz] dependent. μ'_r of a material may increase or decrease with increasing temperature, until a certain temperature (curie temperature) where $\mu'_r = 1$. With increasing signal frequency and increasing current, the value of μ'_r is getting smaller. For example, Mumetal has a μ'_r of over 10'000 at $f = 0$ Hz, but similar to steel at $f = 20$ kHz!

Table G.1 Metals and their properties [1, 2]

	Initial Magnetic Permeability μ_{ri} [1]	Maximum Magnetic Permeability μ_{rm} [1]	Electrical Resistance ϱ [Ωm]	Electrical Conductivity σ [S/m]
Aluminum (Al)	≈1	--	2.62E-08	3.82E+07
Barium (Ba)	≈1	--	--	--
Beryllium (Be)	≈1	--	4.57E-08	2.19E+07
Bismuth (Bi)	≈1	--	1.15E-06	8.70E+05
Brass (66Cu 34Zn)	≈1	--	3.91E-08	2.56E+07
Bronze (4Sn 0.5P, balance Cu)	≈1	--	9.43E-08	1.06E+07
Cadmium (Cd)	≈1	--	7.52E-08	1.33E+07
Chromium (Cr)	≈1	--	2.60E-08	3.85E+07
Cobalt (Co)	60	250	9.71E-08	1.03E+07
Copper (Cu)	≈1	--	1.72E-08	5.82E+07
Ferrite core (MnZn), typical, f<1 MHz	1 000...20 000	--	0.1...25	0,04...10
Ferrite core (NiZn), typical, f<100 MHz	15...300	--	1...20E+05	0,05E-05...1E-05
Gold (Au)	≈1	--	2.13E-08	4.70E+07
Hypernik (50Fe 50Ni)	4’000	70’000		
Iron (99,8Fe), ingot	150	5’000	1.00E-07	1.00E+07
Iron (97Fe 3Si), unoriented	270	8’000	4.69E-07	2.13E+06
Iron (97Fe 3Si), grain oriented	1’400	50’000	5.00E-07	2.00E+06
Lead (Pb)	≈1	--	2.19E-07	4.57E+06
Magnesium (Mg)	≈1	--	4.46E-08	2.24E+07
Metglas 2605SC (81Fe 13B 3,5Si 2C)	--	210’000	1.25E-06	8.00E+05
Metglas 2605S--3 (79Fe 16B 5Si)	--	30’000	1.25E-06	8.00E+05
Mild steel (0,2-0,4C)	100	2’000	1.49E-07	6.70E+06
Mumetal (77Ni 16Fe 5Cu 2Cr)	20’000	350’000	4.69E-07	2.13E+06
Nickel (Ni)	50	1’000	6.90E-08	1.45E+07
Nickel Iron Alloy 4750 (48Ni 52Fe)	11’000	80’000	4.81E-07	2.08E+06
Permalloy 4-79 (79Ni 4Mo 17Fe)	40’000	200’000	5.81E-07	1.72E+06
Permendur (50Fe 50Co)	500	6’000		
Permendur 2V (49Co 51Fe 2V)	800	8’000	4.00E-07	2.50E+06
Platinum (Pt)	≈1	--	1.06E-07	9.40E+06
Sendust (85Fe 10Si 5Al), magn. powder	36’000	120’000	--	--
Silver (Ag)	≈1	--	1.62E-08	6.17E+07
Stainless steel (302, 304)	1.003...5	--	7.09E-07	1.41E+06
Stainless steel (316)	1.003...1.01	--	7.09E-07	1.41E+06
Stainless steel (410)	--	750	5.92E-07	1.69E+06
Stainless steel (430)	--	1’800	5.92E-07	1.69E+06
Steel (SAE 1045)	--	1’000	1.62E-07	6.17E+06
Steel low carbon (99,5Fe)	200	4’000	1.20E-07	8.33E+06
Supermalloy (80Ni 5Mo 15Fe)	80’000	450’000	6.49E-07	1.54E+06
Supermendur (49Co 2V 49Fe)	--	100’000	2.60E-07	3.85E+06
Titanium (Ti)	≈1	--	4.78E-07	2.09E+06
Tin (Sn)	≈1	--	1.14E-07	8.77E+06
Tungsten (W)	≈1	--	5.49E-08	1.82E+07
Waster, distilled	≈1	--	1.00E+04	1.00E-04
Wood, dry	≈1	--	1E+09...1E+13	1E-13...1E-09
Zinc (Zn)	≈1	--	6.10E-08	1.64E+07

G.2 Insulator Properties

To calculate the signal wavelength in a cable or another electric structure surrounded by an insulator (plastics), the *dielectric constant* (dielectric permittivity) ε_r' must be known. Therefore, Tables G.2 and G.3 present the ε_r' of common plastics and other *insulators*. The ε_r' data apply to room temperature of 25 °C (unless otherwise noted).

Table G.2 Plastics and their dielectric constants ε_r' [2]

Material	Dielectric Constant ε_r = Permittivity ε_r [1]			Disspation Factor = Loss Tangent tan(δ) [1]		
	1 kHz	1 MHz	3 GHz	1 kHz	1 MHz	3 GHz
Plastics:						
Alkyd resin	5.1	4.76	4.5	0.0236	0.0149	0.0108
Cellulose acetate-butyrate, plasticized	3.48	3.3	2.91	0.0097	0.018	0.028
Cresylic acid-formaldehyde, 50% α-cellulose	4.95	4.51	3.43	0.033	0.036	0.051
Cross-linked polystyrene	2.59	2.58	2.58	0.0005	0.0016	0.0019
Epoxy resin (Araldite CN-501)	3.67	3.62	3.09	0.0024	0.019	0.027
Epoxy resin (Epon resin RN-48)	3.63	3.52	3.04	0.0038	0.0142	0.021
Foamed polystyrene, 0.25% filler	1.03	1.03	1.03	<0.0001	<0.0002	0.0001
Melamine-formaldehyde, α-cellulose	7.57	7	4.93	0.0122	0.041	0.103
Melamine-formaldehyde, 55% filler	6	5.75	--	0.0119	0.0115	--
Phenol-formaldehyde (Bakelite BM 120)	4.74	4.36	3.7	0.022	0.028	0.0438
Phenol-formaldehyde, 50% paperlaminate	5.15	4.6	3.57	0.0165	0.034	0.06
Polycarbonate (PC)	3.17	3.02	--	0.0021	0.01	--
Polychlorotrifluoroethylene (PCTFE)	2.63	2.42	2.29	0.027	0.0082	0.0028
Polyethylene (PE)	2.26	2.26	2.26	<0.0002	<0.0002	0.00031
Polyethyleneterephthalate (PET)	3.12	2.98	--	0.0047	0.016	--
Polyethylmethacrylate (PEMA)	2.75	2.55	2.51	0.0294	0.009	0.0075
Polyhexamethylene-adipamid (nylon)	3.5	3.14	2.84	0.0186	0.0218	0.0117
Polyimide (PI)	3.5	3.4	--	0.002	0.003	--
Polyisobutylene (PIB)	2.23	2.23	2.23	0.0001	0.0001	0.00047
Polymethyl methacrylate (PMMA)	3.12	2.76	2.6	0.0465	0.014	0.0057
Polyphenylene oxide (PPE)	2.55	2.55	2.55	0.0003	0.0007	0.0011
Polypropylene (PP)	2.25	2.55	--	<0.0005	<0.0005	--
Polystyrene (PS)	2.56	2.56	2.55	<0.00005	0.00007	0.00033
Polytetrafluoroethylene (PTFE, teflon)	2.1	2.1	2.1	<0.0003	<0.0002	0.00015
Polyvinylcyclohexane (PVCH)	2.25	2.25	2.25	0.0002	<0.0002	0.00018
Polyvinyl formal (PVF)	3.12	2.92	2.76	0.01	0.019	0.0113
Polyvinylidene fluoride (PVDF)	8	6.6	--	0.018	0.17	--
Urea-formaldehyde, cellulose	6.2	5.65	4.57	0.024	0.027	0.0555
Urethane elastomer (PUR)	6,7--7,5	6,5--7,1	--	0.055	--	--
Vinylidene-vinyl chloride copolymer (PVDC)	4.65	3.18	2.71	0.063	0.057	0.0072
100% aniline-formaldehyde (Dilecteue-100)	3.68	3.58	3.44	0.0032	0.0061	0.0026
100% phenol-formaldehyde	7.15	5.4	3.64	0.082	0.06	0.052
100% polyvinyl-chloride (PVC)	3.1	2.88	2.84	0.0185	0.016	0.0055

Table G.3 Miscellaneous insulators and their dielectric constants ε_r' [2]

Material	Dielectric Constant ε_r = Permittivity ε_r [1]			Disspation Factor = Loss Tangent tan(δ) [1]		
	1 kHz	1 MHz	3 GHz	1 kHz	1 MHz	3 GHz
Glasses:						
Iron-sealing glass	8.38	8.3	7.99	0.0004	0.0005	0.00199
Soda-borosilicate	4.97	4.84	4.82	0.0055	0.0036	0.0054
100% silicon dioxide (fused quartz)	3.78	3.78	3.78	0.00075	0.0001	0.00006
Rubbers:						
Butyl rubber	2.38	2.35	2.35	0.0035	0.001	0.0009
GR-S rubber	2.96	2.9	2.75	0.0024	0.012	0.0057
Gutta-percha	2.6	2.53	2.4	0.0004	0.0042	0.006
Hevea rubber (pale crepe)	2.4	2.4	2.15	0.0018	0.0018	0.003
Hevea rubber, vulcanized	2.94	2.74	2.36	0.0024	0.0446	0.0047
Neoprene rubber (Polychloroprene)	6.6	6.26	4	0.011	0.038	0.034
Organic polysulfide, fillers	2260	110	16	1.29	0.39	0.22
Silicone-rubber compound	3.35	3.2	3.13	0.0067	0.003	0.0097
Woods:						
Balsa wood	1.4	1.37	1.22	0.004	0.012	0.1
Douglas fir	2	1.93	1.82	0.008	0.026	0.027
Douglas fir, plywood	2.1	1.9	--	0.0105	0.023	--
Mahogany	2.4	2.25	1.88	0.012	0.025	0.025
Yellow birch	2.88	2.7	2.13	0.009	0.029	0.033
Yellow poplar	1.79	1.75	1.5	0.0054	0.019	0.015
Miscellaneous:						
Amber (fossil resin)	2.7	2.65	2.6	0.0018	0.0056	0.009
DeKhotinsky cement	3.75	3.23	2.96	0.0335	0.024	0.021
Gilsonite (99.9% natural bitumen)	2.66	2.58	--	0.0035	0.0016	--
Shellac (natural XL)	3.81	3.47	2.86	0.0074	0.031	0.0254
Mica, glass-bonded	7.45	7.39	--	0.0019	0.0013	--
Mica, glass, titanium dioxide	9.3	9	--	0.125	0.0026	0.004
Ruby mica	5.4	5.4	5.4	0.0006	0.0003	0.0003
Paper, royal grey	3.29	2.99	2.7	0.0077	0.038	0.056
Soil, sandy dry	2.91	2.59	2.55	0.08	0.017	0.0062
Soil, loamy dry	2.83	2.53	2.44	0.05	0.018	0.0011
Ice (from pure distilled water, −12 °C)	--	4.15	3.2	--	0.12	0.0009
Freshly fallen snow (−20 °C)	3.33	1.2	1.2	0.492	0.0215	0.00029
Hard-packed snow followed by light rain (−6 °C)	--	1.55	1.5	--	0.29	0.0009
Water (distilled)	--	78.2	76.7	--	0.04	0.157

G.3 PCB Materials

The dielectric constant (relative permittivity ε_r') of a *printed circuit board* (PCB) base material determines the characteristic impedance of the PCB traces and is, therefore, an important parameter in the field of EMC and signal integrity. Besides the dielectric constant ε_r', the loss tangent $\tan(\delta)$ (also called dissipation factor Df) of a PCB material is also of interest because it influences the loss at high frequencies (together with the ohmic loss, which increases with increasing frequency due to the skin effect).

The arrangement of the copper and insulation layers of a PCB is called the PCB layer stackup, or just stackup. Let us have a look at an example of a PCB stackup

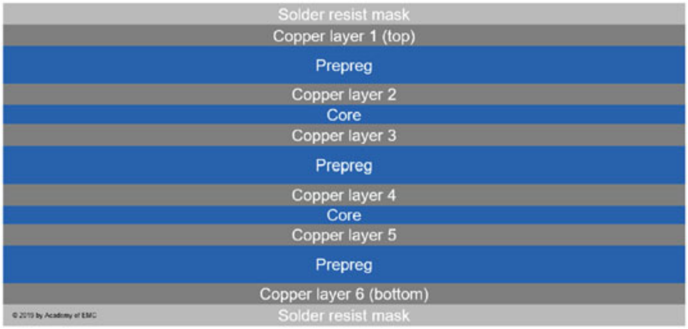

Fig. G.2 Example of a PCB stackup with layers

of a six-layer board in Fig. G.2. These are the following materials that one must consider when calculating the characteristic impedance of PCB traces:

- **Solder resist mask.** The solder mask is a thin (usually green) layer that protects the copper conductors from oxidation and mechanical stress and helps minimize the creation of short circuits through bridges formed by excess solder. The typical thickness of the solder mask (above the copper conductors) is 0.8 mils = 20 μ m. The dissipation factor (loss tangents) is usually 0.025 @ 1 GHz and the dielectric constant 3.3 to 3.8.
- **Copper layer.** The copper layers consist of thin, rolled and annealed (RA), or electro-deposited (ED) copper. Typical copper layer thickness:
  - 0.5 oz = 0.7 mils = 17.5 μ m
  - 1 oz = 1.4 mils = 35 μ m
  - 2 oz = 2.8 mils = 70 μ m

  The circuit traces are etched into the copper layers before the PCB is laminated together (with adhesive, heat, and pressure).
- **Core.** PCB cores are *laminates* (PCB base materials) with copper layers on both sides. The circuit traces are etched into their copper layers before the cooper clad laminates are glued together with the rest of the multilayer PCB. The distance between the two copper layers of a core has only slight variation and the characteristic impedance can be controlled with high accuracy. Cores are typically made out of *FR-4* substrates because FR-4 has a good price-to-performance ratio regarding low dissipation factor, low variation of ε_r' over a wide frequency range, and moisture absorption.
- **Prepreg.** Prepreg is the short word for pre-impregnated. It is a flexible material, typically also containing woven glass, which is supplied to the PCB fabricator partially cured (not completely cooked). It is included between the rigid core layers in the layer stack during fabrication and then heated to perform final curing, after which it becomes rigid, helping to join the core substrates of the finished PCB. For boards that require four or more layers, core and prepreg layers

are interleaved to build up the required number of layers. The cores are all etched individually and then sandwiched together with layers of prepreg on the top and bottom and bonding the two cores together. Cores are typically made out of FR-4 substrates because FR-4 has a good price-to-performance ratio regarding low dissipation factor, low variation of ε_r' over a wide frequency range, and moisture absorption.

Here are some tips regarding PCB layer stackups:

1. **Power decoupling planes.** Design a PCB stackup with power supply plane and GND plane close together (< 0.1 mm or 2...3 mils). This leads to an especially good decoupling at high frequencies (> 1 MHz).
2. **Impedance controlled routing.** Generally, core is more reproducible than prepreg regarding thickness and dielectric constant. This means that controlled impedance layers should ideally be routed along the core material, rather than prepreg.
3. **FR-4.** FR-4 is not equal FR-4. However, if the exact PCB data are unknown, $\varepsilon_r' = 4.5$ and $\tan(\delta) = 0.015$ can be assumed for FR-4.

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Appendix H
Galvanic Series

Learn what is to be taken seriously and laugh at the rest.

—Hermann Hesse

Knowing the galvanic series is important when designing interconnections between dissimilar metals to bond shields or safety earth connections. If you do not care which metals you use, the interconnections may corrode within a short time, and the shielding effect or safety function is not given anymore.

H.1 Corrosion

Why does *corrosion* happen and recommendations for connections of dissimilar metals:

- **Corrosion.** The less noble metal (anode) of an interconnection of two metals experiences galvanic corrosion if:
  1. The two metals have a galvanic incompatibility (voltage difference to high). A difference of hundreds of millivolts is likely to result in galvanic corrosion, but only a few tens of millivolts is unlikely to be a problem.
  2. An electrolyte (e.g., water, moisture) is present.
  3. The two metals have an electrical conducting connection.
- **Rate of corrosion.** The rate of corrosion depends on the moisture of the environment, how far apart the metals in the galvanic series are (further apart leads to faster corrosion, because the ion transfer is faster), and other parameters like the type of electrolyte (pH, concentration, flow rate), temperature (rate of corrosion increased with higher temperatures), degree of aeration, humidity, pressure, and even the geometry of the interconnection.

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
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- **Recommendations.** The galvanic series helps us choose the right combination of metals for an interconnection in terms of corrosion. Here is a rule-of-thumb on how to choose metals depending on their potential difference with respect to environment [4]:
  - **Harsh** (outdoor, high humidity, salt laden, military). Choose metals where electrode potential difference is ≤ 0.15 V.
  - **Normal** (non-temperature/humidity controlled, consumer product, indoor). Choose metals where electrode potential difference is ≤ 0.3 V.
  - **Controlled** (temperature/humidity controlled, indoor). Choose metals where electrode potential difference is ≤ 0.5 V.

H.2 Corrosion Prevention

Three methods for corrosion prevention of dissimilar metal connections:

- **Isolation.** If no electrically conductive connection is necessary, put an insulator between the dissimilar metals to avoid direct contact.
- **Low potential difference.** If an electrically conductive connection between dissimilar metals is a must, the dissimilar metals should have a low voltage difference in the galvanic series.
- **Keep interconnection dry.** Another option for preventing corrosion is to keep the interconnection between the dissimilar metals dry, e.g., with a coating (no electrolyte = no corrosion).

H.3 Standard Electrode Potential Series

Table H.1 shows *standard electrode potentials* (E°) for various metals. The standard electrode potential is defined by measuring the potential relative to a standard hydrogen electrode (SHE) using 1 mol solution at 25 °C at the pressure of 1 atm.

An electrode potential series can be derived for metals in any electrolyte solution. Be careful! The real-world corrosion rate depends on the solution conditions like electrolyte concentration, pH, flow rate, aeration, temperature, humidity, and pressure. Therefore, it is common to use the seawater electrode potential Table H.2 rather than the standard hydrogen electrode potential Table H.1.

Table H.1 Standard electrode potentials [5]

Standard electrode potentials, aqueous solution (1mol ion / 1l), respective metal ion, 25 ºC, at equilibriumo [V]	**Top = least noble** (anodic) ↓ **Bottom = most noble** (cathodic)
–3.02	Lithium* (Li)
–2.92	Cesium* (Cs)
–2.92	Potassium* (K)
–2.84	Calcium^{2+} (Ca)
–2.71	Sodium* (Na)
–2.35	Magnesium^{2+} (Mg)
–1.67	Aluminum^{3+} (Al)
–1.05	Manganese^{2+} (Mg)
–0.76	Zinc^{2+} (Zn)
–0.74	Chromium^{3+} (Cr)
–0.44	Iron^{2+} (Fe)
–0.40	Cadmium^{2+} (Cd)
–0.28	Cobalt^{2+} (Co)
–0.24	Nickel^{2+} (Ni)
–0.14	Tin^{2+} (Sn)
–0.13	Lead^{2+} (Pb)
0.00	Hydrogen* (H)
0.35	Copper^{2+} (Cu)
0.52	Copper^{3+} (Cu)
0.80	Silver* (Ag)
0.86	Quicksilver^{2+} (Hg)
1.20	Platinum^{2+} (Pg)
1.42	Gold^{3+} (Au)

* Superscripted symbols refer to the electrochemical valence.

H.4 Seawater Galvanic Series

The seawater *galvanic series* is often used to approximate the probable galvanic effects in other environments for which there are no data. For example, from the standard electrode potentials shown in Table H.1, it can be seen that aluminum (Al) should behave anodically toward zinc (Zn) and presumably would retard the corrosion of zinc in a usual coupled situation. However, the reverse is true, as can be seen from the established galvanic series of metals in seawater in Table H.2.

In Table H.2, metals are grouped. All metals, alloys, and platings of the same group have similar electro-motive forces (EMF) within 0.05 V when coupled with a saturated calomel electrode in seawater at room temperature. All members of a group, regardless of metallurgical similarity or dissimilarity, are considered

Table H.2 Seawater electrode potentials [1, 2]

Group Nr.	Metallurgical Category	EMF [V]	Compatible Couples
1	**Gold (Au)**: solid and plated Goldplatinum alloys Wrought **platinum (Pt)**	+0.15	
2	**Rhodium (Rh)** plated on silverplated copper	+0.05	
3	**Silver (Ag)**: solid or plated High silver alloys	0	
4	**Nickel (Ni)**: solid or plated Monel metal High-nickel-copper (CuNi) alloys	−0.15	
5	**Copper (Cu)**: solid or plated Low **brasses** or **bronzes** Silver (Ag) solder German silvery high copper-nickel (CuNi) alloys **Nickelchromium (NiCr)** alloys Austenitic corrosion-resistant **steels**	−0.20	
6	Commercial **yellow brasses** and **bronzes**	−0.25	
7	**High brasses** and **bronzes** Naval brass Muntz metal	−0.30	
8	18% chromium (Cr) type corrosion-resistant steels	−0.35	
9	**Chromium (Cr)** plated **Tin (Sn)** plated 12% chromium type corrosion resistant steels	−0.45	
10	**Tin (Sn)** plate Terne-plate Tin lead (SnPb) solder	−0.50	
11	**Lead (Pb)**: solid or plated High lead (Pb) alloys	−0.55	
12	**Aluminum (Al)** wrought alloys of the 2000 Series	−0.60	
13	**Iron (Fe)**, wrought, gray or malleable Plain **carbon steel** and low alloy steels Armco iron	−0.70	
14	**Aluminum (Al)**: wrought alloys other than 2000 Series **Aluminum (Al)**: cast alloys of the silicon type	−0.75	
15	**Aluminum (Al)**: cast alloys other than silicon type **Cadmium (Cd)**: plated and chromated (Cr)	−0.80	
16	Hot-dip-zinc (Zn) plate **Gvanized steel**	−1.05	
17	**Zinc (Zn)**: wrought Zinc (Zn) base diecasting alloys Zinc (Zn) plated	−1.10	
18	**Magnesium (Mg)** & magnesiumbase alloys: cast or wrought	−1.60	

Indicates the most cathodic (most noble) members of the series
Indicates an anodic member (least noble) of the series

compatible. Compatible couples between groups have been specified in the table (green areas) based on a potential difference of 0.25 V maximum.

In Table H.3, the galvanic series of selected metals in seawater is presented. This series can be used as a reference to minimize galvanic corrosion when selecting metals that will be in direct contact. Generally said, the closer the metals in the series, the less galvanic corrosion is expected. In a galvanic couple, the metal higher in the series represents the anode and will corrode preferentially in the environment to the cathode, which is lower in the series.

Table H.3 Galvanic series of selected metals in seawater [3]

Least noble (Anodic)	Iron, cast	Stainless steel 201 (active)
	Stainless steel 410 (active)	Carpenter 20 (active)
Magnesium	Copper (plated, cast or wrought)	Stainless steel 321 (active)
Mg Alloy AZ-31B	Nickel (plated)	Stainless steel 316 (active)
MG Alloy HK-31A	Chromium (plated)	Stainless steel 309 (passive)
Zinc (hot-dip, die cast or plated)	Tantalum	Stainless steel 17-7 PH (passive)
Beryllium (hot pressed)	AM350 (active)	Silicone Bronze 655
Aluminum (Al) 7072 cl. on 7075	Stainless steel 310 (active)	Stainless steel 304 (passive)
Al alloy 2014-T3	Stainless steel 301 (active)	Stainless steel 301 (passive)
Al alloy 1160-1114	Stainless steel 304 (active)	Stainless steel 321 (passive)
Al alloy 7079-T6	Stainless steel 430 (passive)	Stainless steel 201 (passive)
Cadmium (plated)	Stainless steel 410 (passive)	Stainless steel 286 (active)
Uranium	Stainless steel 17-7 PH (active)	Stainless steel 316L (passive)
Al alloy 218 (die cast)	Tungsten	AM355 (active)
Al alloy 5052-0	Niobium (Columbium) 1% Zr	Stainless steel 202 (passive)
Al alloy 5052-H12	Brass, yellow, 268	Carpenter 20 (passive)
Al alloy 5456-0, H353	Uranium 8% Mo	AM355 (passive)
Al alloy 5052-H32	Brass, Naval, 464	A286 (passive)
Al alloy 1100-0	Yellow brass	Titanium 5A1,2.5Sn
Al alloy 3003-H25	Muntz metal 280	Titanium 13V,11Cr,3A1 (annealed)
Al alloy 6061-T6	Brass plated	Titanium 6A1,4V (solution treated + aged)
Al alloy A360 (die cast)	Nickel-silver (18% Ni)	Titanium 6A1,4V (annealed)
Al alloy 7075-T6	Stainless steel 316L (active)	Titanium 8 Mn
Al alloy 1160-1114	Bronze 220	Titanium 13V,11Cr,3A1 (solution treated + aged)
Al alloy 6061-0	Copper 110	Titanium 75A
Indium	Red brass	AM350 (passive)
Al alloy 2014-0	Stainless steel 347 (active)	AM350 (passive)
Al alloy 2024-T4	Molybdenum, commercial pure	Silver
Al alloy 5052-H16	Copper-Nickel 715	Gold
Tin (plated)	Admiralty brass	Graphite
Stainless steel 430 (active)	Stainless steel 202 (active)	
Lead	Bronze, Phosphor 534 (B-1)	**Most noble** (Cathodic)
Steel 1010	Monel 400	

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Appendix I
Triboelectric Effect

What we know is a drop, what we don't know is an ocean.

—Isaac Newton

The *triboelectric effect* is the physical effect that leads to electrically charged materials after they are separated from a different material with which they were in contact. Rubbing two materials with each other increases the contact between their surfaces, hence the triboelectric effect.

Most everyday static electricity is triboelectric. The polarity and strength of the charges produced differ according to the materials, surface roughness, temperature, humidity, strain, and other properties. These static charges are why electrical apparatus and installations have to be tested for their immunity to *electrostatic discharges* (ESD).

I.1 How Electrostatic Charges Are Produced

When two or more non-conductive materials with different affinities for electrons make and break contact, electrostatic voltage potentials up to several ten thousand volts may result. The reason why this happens is that electrons can be exchanged between materials on contact. Materials with weakly bound electrons tend to lose them, while materials with sparsely filled outer shells tend to gain them. Every time a material with weakly bound electrons and a material with the tendency to gain electrons make contact, electrons spend more time on the material with the higher affinity for electrons. When conductive materials are separated, electrons quickly return to the material they initially came from. However, when the materials are not conductive, the separation of the materials may occur before the electrons are back to the material where they initially came from. This electron exchange leads to a positively charged material (the one that lost electrons) and a negatively charged material.

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In practice, the magnitude of charge depends heavily on the relative humidity rH [%] of air. In a humid environment, the slight amount of moisture on the surface of a non-conductive material can lead to a rise of conductivity high enough to prevent significant charge separation. Therefore, the larger the electrical charges are, the lower the relative humidity. At an optimal relative humidity of 50 %–55 %, the conductivity of the material surfaces is increased to such an extent that electrical charges can be dissipated without any problems [1]. However, bear in mind that the often prevailing opinion that moist air conducts electricity better than dry air does not apply. It is the moisture on the surface of materials, which prevents static electricity.

I.2 Triboelectric Series

The triboelectric series is a list of materials ranking from the most positive (materials that tend to lose electrons, Fig. I.1) to the most negative (materials that tend to gain electrons).

Fig. I.1 Triboelectric series [2]

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Appendix J
Country-Specific EMC Regulations

> *If a product cannot be legally sold because of its inability to comply with the governmental EMC requirements, the fact that it may perform a function that gives it significant sales potential is unimportant!*
>
> —Dr. Clayton R. Paul

Every government issues its own EMC regulations (laws, directives) for its country. This chapter lists the legal statutory basis and requirements regarding EMC for different countries and customs unions.

The author can accept no responsibility or liability for the content of this chapter and throughout this book. All information provided in this chapter and throughout the book are provided as is, with no warranties, and should not be used as a reference for compliance testing. Please make sure to test your product according to the recognized EMC standards at an accredited EMC laboratory.

J.1 Australia and New Zealand

The Regulatory Compliance Mark (RCM, Fig. J.1) is the compliance mark for Australia and New Zealand. By the end of March 2016, the RCM replaced the former compliance marks called A-Tick (telecommunications equipment requirements) and C-Tick (EMC requirements).

- **Effective countries.** Australia, New Zealand.
- **Responsible authority.** EMC: Australian Communications and Media Authority (ACMA, Australia), Radio Spectrum Management Group (RSM, New Zealand). Electrical Safety: Electrical Regulatory Authorities Council (ERAC).
- **Statutory basis.** EMC regulatory arrangements (Radiocommunications Act 1992).
- **Scope of RCM marking.** All applicable ACMA regulatory arrangements. This includes telecommunications, radio communications, electromagnetic compati-

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Fig. J.1 Regulatory Compliance Mark (RCM). The mandatory compliance mark for Australia and New Zealand

Fig. J.2 ANATEL label is mandatory for radio- and telecommunications equipment in Brazil

bility (EMC), electromagnetic energy (EME), electromagnetic radiation (EMR), electrical safety [13].

- **RCM marking obligation for electronic products.** Mandatory [14].
- **Emissions tests required.** Yes [5].
- **Immunity tests required.** No [5].
- **EMC standards.** Australian and New Zealand EMC standards (AS/NZS), which are based on IEC and CISPR standards. ACMA mandated EMC standards are published online [12].

J.2 Brazil

The ANATEL label (Fig. J.2) is mandatory for radio- and telecommunications equipment in Brazil. It consists of the ANATEL logo and a numerical code HHHHH-AA-FFFFF [36], where:

- HHHHH identifies the product (5-character sequential numbering).
- AA identifies the year of issue of the approval (2-digit type).
- FFFFF identifies the product manufacturer (5 numeric characters).

- **Effective countries.** Brazil.
- **Responsible authority.** Agência Nacional de Telecomunicações.
- **Statutory basis.** Resolution about certification and approval of telecommunication products: Resolution No. 715 (23 October 2019) [39]. Resolution about numerical code: Resolution No. 662 (8 March 2016) [38].
- **Scope of ANATEL label.** Telecommunication products (see Ato No. 2222 (20 April 2020)) [35].
- **ANATEL marking obligation for electronic products.** For telecommunication products [35].

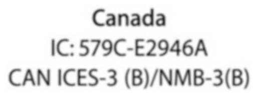

Fig. J.3 The ISED label of an Apple iPhone 6s

- **Emissions tests required.** Yes [34].
- **Immunity tests required.** Yes [34].
- **EMC standards.** ANATEL publishes acts (called Ato) which contain the applicable limits and refer to the applicable EMC and safety standards (IEC/CISPR/ITU-T standards). For EMC, the act Ato No. 1120 is valid (former resolution 442) [35]. For safety, the act Ato No. 950 is valid (former resolution 529, basically IEC 60950 (2005)) [37].

J.3 Canada

The Innovation, Science and Economic Development (ISED) certification number is made up of a Company Number (CN) assigned by ISED's CEB, followed by the Unique Product Number (UPN). For example, IC: CN-UPN. The CN can only be obtained from the Certification and Engineering Bureau (CEB) of ISED. The registration number (Fig. J.3) can be presented electronically (on the device's display) or physically on a product label. An FCC test report will be accepted by ISED if it is less than 1 year old [27].

- **Effective countries.** Canada.
- **Responsible authority.** Innovation, Science and Economic Development Canada (ISED, equivalent to United States' FCC) [26]. Certification and Engineering Bureau (provides a certification service for radio equipment and a registration service for terminal equipment) [25].
- **Statutory basis.** Radiocommunication Act (R.S.C., 1985, c. R-2).
- **Product label.** ISED certification number (preceded with "IC:"), mandatory.
- **Emissions tests required.** Yes (FCC test reports are accepted [27]).
- **Immunity tests required.** No [27].
- **EMC standards.** Emission limits: Interference-Causing Equipment Standards (ICES) refer to CAN/CSA-CISPR standards [8]. Test methods: CISPR 16 [8], ANSI C63.x [9].

J.4 China

The China Compulsory Certificate mark (CCC) is a mandatory mark for products sold on the Chinese market (Fig. J.4).

- **Effective countries.** China.

Fig. J.4 The China Compulsory Certificate mark (CCC)

Fig. J.5 The Eurasian Conformity mark (EAC)

- **Responsible authority.** State Administration for Market Regulation (SAMR) [18].
- **Scope of CCC mark.** Safety, health, environmental protection, electromagnetic compatibility (emission and immunity) and other technical regulations or standard.
- **Statutory basis.** Regulations for Compulsory Product Certification.
- **CCC marking obligation for electronic products.** Mandatory.
- **EMC standards.** GB standards (GB stands for Guobiao, which means National Standard). GB standards are based on the international IEC standards.

J.5 Eurasian Economic Union

The Eurasian Conformity (EAC) mark (Fig. J.5) is the product conformity mark of the Eurasian Economic Union (EAEU, former Eurasian Customs Union (EACU)).

- **Effective countries.** Armenia, Belarus, Kazakhstan, Kyrgyzstan, and Russia (Eurasian Economic Union) [41].
- **Responsible authority.** Eurasian Economic Commission (EEC).
- **Statutory basis.** Technical Regulations of the Custom Union electromagnetic compatibility of technical devices: CU TR 020/2011 [33].
- **Scope of EAC.** Safety, health, environmental protection, electromagnetic compatibility, and other technical regulations or standard.
- **EAC marking obligation for electronic products.** Mandatory.
- **Emissions tests required.** Yes [33].
- **Immunity tests required.** Yes [33].
- **EMC standards.** Russian GOST-R standards, which are mostly identical to the European EN standards.

Fig. J.6 The CE conformity mark

J.6 European Union

The Conformité Européenne (CE) mark (Fig. J.6) is the product conformity mark of the European Union (EU) [11].

- **Effective countries.** European Economic Area (EEA). 28 EU Member States, Norway, Iceland and Liechtenstein (the so-called EEA-EFTA States), and Switzerland and Turkey.
- **Responsible authority.** European Union.
- **Statutory basis.**
  - **EMC Directive (EMCD).** Directive 2014/30/EU [42] of the European Parliament and of the Council of 26 February 2014 on the harmonization of the laws of the Member States relating to electromagnetic compatibility. Be aware that the EMC directive can be overruled by other EU directives or regulations [2014/30/EU article 2, paragraph 3]. Examples are:

    Automotive. Regulation (EC) 661/2009. The Guide for the EMCD [21] states that vehicles and equipment subject to type approval under UNECE Regulation 10 are totally excluded from the scope of the EMCD. The Commission Regulation (EU) 2019/543 of 3 April 2019 amending Annex IV to Regulation (EC) 661/2009 refers for electromagnetic compatibility to the United Nations Economic Commission for Europe Regulation 10 (UNECE R 10).
    Radio equipment. Directive 2014/53/EU. The RED overrules the EMCD according to the introductory "whereas"-text, point (8). However, in article 3, paragraph 1b, the RED calls for the same protection regarding EMC as the EMCD. Generally spoken, regarding EMC testing for equipment under RED, the same harmonized EMC standards—as for EMCD—plus the ETSI EN 301 489-x series have to be applied.
    Medical. Regulation (EU) 2017/745. The Medical Device Regulation (MDR) overrules the EMCD with its article 1, paragraph 11.
    In vitro diagnostics. Regulation (EU) 2017/746. The In vitro Diagnostics Regulation (IVDR) overrules the EMCD with its article 1, paragraph 5.
  - **CE marking.** Regulation (EC) 765/2008 for accreditation and market surveillance (RAMS).
- **Scope of CE marking.** Safety, health, environmental protection, electromagnetic compatibility, and other technical regulations or standard.
- **CE marking obligation for electronic products.** Mandatory. However, not all products must have CE marking. It is compulsory only for most of the products

Emission [test methods]:

- CISPR 11 → EN 55011 - Emission limits and methods for ISM equipment
- CISPR 32 → EN 55032 - Emission limits and methods for multimedia equipment
- CISPR 14 → EN55014-1 - Emission limits and methods for household appliances
- IEC 61000-3-2 → EN 61000-3-2 - Methods for harmonic current emissions
- IEC 61000-3-3 → EN 61000-3-3 - Methods for voltage fluctuation and flicker

Immunity [test methods]:

- IEC 61000-4-2 → EN 61000-4-2 - Test methods for electrostatic discharge (ESD)
- IEC 61000-4-3 → EN 61000-4-3 - Test methods for radiated RF field immunity
- IEC 61000-4-4 → EN 61000-4-4 - Test methods for burst (EFT)
- IEC 61000-4-5 → EN 61000-4-5 - Test methods for mains surge
- IEC 61000-4-6 → EN 61000-4-6 - Test method for conducted RF immunity
- IEC 61000-4-8 → EN 61000-4-8 - Test methods for mains frequency magnetic field
- IEC 61000-4-11 → EN 61000-4-11 - Test methods for AC supply dips, interruptions
- IEC 61000-4-39 → EN 61000-4-39 - Test methods for radiated close proximity fields

Generic standards [emission limits, immunity levels]:

- IEC 61000-6-3 → EN 61000-6-3 - Emission limits for residential environments
- IEC 61000-6-4 → EN 61000-6-4 - Emission limits for industrial environments
- IEC 61000-6-8 → EN 61000-6-8 - Emission limits for commercial, light-industrial
- IEC 61000-6-1 → EN 61000-6-1 - Immunity levels for residential, light-industrial
- IEC 61000-6-2 → EN 61000-6-2 - Immunity levels for industrial environments

Fig. J.7 Some IEC/CISPR standards and their corresponding EN counterparts

covered by the New Approach Directives. It is forbidden to affix the CE marking to products not covered by at least one of the New Approach Directives [1].

- **Emissions tests required.** Yes [42].
- **Immunity tests required.** Yes [42].
- **EMC standards.** European standards, including *Harmonized Standards*, are often based fully or partially on international ISO or IEC standards and they start with the letters EN, which stands for European Norm. EN EMC standards are developed by CENELEC, CEN, and ETSI. Figure J.7 shows some IEC/CISPR standards and their corresponding EN counterparts.

  An important concept of EMC standards and product compliance within the EU is the presumption of conformity. Products that proved compliance with the Harmonized Standards (published via EU Official Journal (OJEU) [2]) benefit from a presumption of conformity with the corresponding essential requirements (directives and regulations) of the applicable legislation [40]. Harmonized Standards are European standards adopted upon a request made by the European Commission to apply EU harmonization legislation.

  Application of harmonized or other standards is voluntary. The manufacturer can always apply other technical specifications to meet the requirements (but will carry the burden of demonstrating that these technical specifications answer the needs of the essential requirements, more often than not, through, a process involving a third-party conformity assessment body—in other words, reversal of the burden of proof) [40].

  - European Standardisation Organisations (EOS = CEN, CENELEC, ETSI) publish standards and their *Date Of Withdrawal* (DOW) [10].
  - The EU publishes (via its Official Journal (OJEU) [2]) which harmonized EMC standards have to be applied for presumption of conformity. Important: the DOW of an EMC standard (published, e.g., by CENELEC) is not legally binding. The regulatory/legal date where an EMC standard losses its status of *presumption of conformity* is the *Date Of Cessation* (DOC), published by the EU via OJEU.
  - The EU publishes harmonized standards for every EU directive online [1].

  Besides the Harmonized EMC Standards, there are is also sector-specific EMC standards like the ED-14D by the European Organization for Civil Aviation Equipment (EUROCAE). The ED-14D standard defines the Environmental Conditions and Test Procedures for Airborne Equipment. ED-14D is harmonized with the US RTCA DO-160 standard [32].
- **Guidance.**
  - **EMCD.** Check the EMCD Guide [21] if your product falls under the EMCD and if you need a CE marking. Moreover, the Radio Equipment Directive Compliance Association (REDCA) published freely available Technical Guidance Notes (TGN) reagarding the EMCD [3].
  - **RED.** The Radio Equipment Directive Compliance Association (REDCA) published the EUANB Technical Guidance Notes (TGN) regarding the RED [4].

J.7 India

The Indian Standards Institution (ISI) mark (Fig. J.8) is the product conformity mark of India, and the Telecommunication Engineering Centre (TEC) logo is mandatory for telecommunication equipment.

- **Effective countries.** India.
- **Responsible authority.** Ministry of Communication and Information Technology and in addition for telecommunication equipment the Department of Telecommunications.

Fig. J.8 Left: Telecommunication Engineering Centre (TEC) logo. Right: Indian Standards Institution (ISI) mark

- **Statutory basis.** Bureau of Indian Standards Act 2016 No. 11., Indian Telegraph Amendment Rules 2017 (TEC logo, Mandatory Testing and Certification of Telecom Equipment (MTCTE) certification).
- **Scope of ISI marking.** Safety, health, environmental protection, electromagnetic compatibility, and other technical regulations or standard.
- **Scope of TEC logo (MTCTE certification).** Performance of telecommunication network, safety of end-users, and radio-frequency emissions.
- **ISI marking obligation for electronic products.** The list of products under mandatory certification can be found online [23].
- **TEC logo obligation for electronic products.** The MTCTE certification and therefore the TEC logo is mandatory for telecommunication equipment [22].
- **EMC standards.** Indian Standards (IS) developed by the Bureau of Indian Standards (BIS), which are based on IEC and CISPR standards [24]. For the Mandatory Testing and Certification of Telecom Equipment (MTCTE), the testing and certification procedure TEC 93009 defines the applicable TEC standards and the TEC standards define the applicable IEC or ETSI standards.

J.8 Japan

The PSE Category A label (diamond) is the Japanese conformity marking for Specific Products. The PSE Category B label (round) is the Japanese conformity mark for Non-Specific Products. PSE stands for Product Safety Electrical Appliance & Material. The Ministry of Economy, Trade and Industry (METI) decides which products belong to Category A (specified products) and which products belong to Category B (non-specified products). The difference for Category A (diamond) to Category B (round) is that Category A products must pass an additional conformity assessment by a Registered Conformity Assessment Body [28].

The Voluntary Control Council for Interference (VCCI) marking is the voluntary Japanese conformity marking for Information Technology Equipment (ITE) based on the CISPR recommendations (Fig. J.9).

- **Effective countries.** Japan.

Fig. J.9 The Japanese compliance labels for Specified Electrical Appliances and Materials (Category A, diamond), Non-Specified Electrical Appliances and Materials (Category B, round), and the Voluntary Control Council for Interference marking (VCCI)

Fig. J.10 The Korean Certification (KC) mark

- **Responsible authority.** Ministry of Economy, Trade and Industry (METI).
- **Statutory basis.** Electrical Appliances and Materials Safety Act (DEN-AN-law).
- **Scope of labels.** Safety (PSE). EMC and EMI (VCCI).
- **PSE and VCCI labeling obligation for electronic products.**
  - **PSE Category A (diamond).** Mandatory for Specified Products [28].
  - **PSE Category B (round).** Mandatory for Non-Specified Products [28].
  - **VCCI.** Voluntary control of radio disturbances emitted from Information Technology Equipment (ITE).
- **EMC standards.** JIS standards which are based on CISPR and IEC standards.

J.9 Republic of Korea

The Korean Certification (KC) marking (Fig. J.10) is the product conformity mark of the Republic of Korea (South Korea).

- **Effective countries.** Republic of Korea (South Korea).
- **Responsible authority.** EMC: Radio Research Agency (RRA). Safety: Korean Agency for Technology and Standards (KATS).
- **Statutory basis.** EMC: Radio Wave Law. Safety: Electric Appliances Safety Control Act.
- **Scope of KC marking.** Safety, health and EMC.
- **KC marking obligation for electronic products.** Mandatory.
- **EMC standards.** Korean EMC Standards (KS), which are mostly identical to the International Standards (IEC). See, e.g., annex 6 of the Free Trade Agreement (FTA) EU-Korea FTA User Guide for Electromagnetic Compatibility (EMC) and Electric Safety Certification [16].

J.10 Switzerland

The CH mark is the Swiss conformity mark. Switzerland also accepts products with the CE mark (according to the agreement between the European Community and the Swiss Confederation on mutual recognition in conformity assessment [6]) (Fig. J.11).

- **Effective countries.** Switzerland.
- **Responsible authority.** Federal Office Of Communications (OCOM, BAKOM).

Fig. J.11 The Swiss conformity (CH) mark

Fig. J.12 The BSMI mark is the conformity mark of Taiwan

- **Statutory basis.** 734.5 Ordinance of 25 November 2015 on Electromagnetic Compatibility (OEMC).
- **Scope of CH marking.** Electromagnetic Interference (EMI), Electromagnetic Compatibility (EMC).
- **CH marking obligation for electronic products.** Either CH or CE must be placed on product. CE is not mandatory (because the CE mark belongs to the EU).
- **Emissions tests required.** Yes [29].
- **Immunity tests required.** Yes [29].
- **EMC standards.** OCOM releases EMC standards. Basically, they are identical to the EN standards.

J.11 Taiwan

The Commodity Inspection Mark (Fig. J.12), also named Bureau of Standards, Metrology and Inspection (BSMI) Mark or Product Safety Mark, is the product conformity mark of Taiwan for Information Technology Equipment (ITE), audio and video devices, appliances, and wireless products.

- **Effective countries.** Taiwan.
- **Responsible authority.** Bureau of Standards, Metrology and Inspection (BSMI).
- **Statutory basis.** Commodity Inspection Act and the respective regulations and directions.
- **Scope of BSMI mark.** Safety, health, environmental protection, electromagnetic compatibility and other technical regulations or standard [7].
- **BSMI marking obligation for electronic products.** Mandatory.
- **EMC standards.** Most of the Chinese National Standards (CNS) are derived from IEC and CISPR EMC standards.

Fig. J.13 The TSE Mark is the Turkish conformity mark

J.12 Turkey

According to the EMCD Guide about the EMC Directive (EMCD) 2014/30/EU [21], the EMCD also applies to Turkey. Therefore, Turkey accepts the CE mark if the declaration of conformity for the CE mark is translated into Turkish.

For radio equipment, the used radio frequency determines how to get the approval. If the equipment uses harmonized frequencies (EU, Turkey), the RED 2014/53/EU (2014/53/AB) can be applied. In the case of non-harmonized frequencies (frequencies that are not allowed or regulated in EU), a Turkish BTK type approval is required.

The TSE Mark is the Turkish conformity mark (Fig. J.13) for electronic products, indicating compliance of a product to relevant Turkish Standards (TSE) and permits using the TSE mark upon the product and/or packaging. In case a product has the CE conformity, the TSE mark is optional.

The TSEK Mark (Certificate of Conformity to the TSE Criteria) is the Turkish conformity mark if there are no relevant Turkish Standards available and your product is not CE certified. Such products are subject to the technical conditions accepted by the Turkish Standards Institute based on the conditions of Turkey and/or other countries.

- **Effective countries.** Turkey
- **Responsible authority.** Ministry of Science, Industry and Technology (SANAYI), Information and Communication Technologies Authority (BTK).
- **Statutory basis.**
  - EMC Directive (EMCD). ELEKTROMANYETİK UYUMLULUK YÖNETMELİĞİ (2014/30/AB).
  - Radio Equipment Directive (RED). TELSİZ EKİPMANLARI YÖNETMELİĞİ TASLAĞI (2014/53/AB).
- **Emissions tests required.** Yes [15].
- **Immunity tests required.** Yes [15].
- **EMC standards.** The Turkish Standards Institution (TSE) releases the EMC standards for Turkey. Basically, they are identical to the EN standards.

Fig. J.14 The United Kingdom Conformity mark (UKCA mark) is the mandatory conformance mark in Great Britain

J.13 United Kingdom

The United Kingdom (UK) left the EU on 1 February 2020. The UKCA mark (Fig. J.14) is the new mandatory conformity mark in Great Britain (England, Wales, and Scotland) to indicate conformity of a product to the UK legislation.

Special rules apply for Northern Ireland. The Northern Ireland Protocol came into force on 1 January 2021. For as long as it is in force, Northern Ireland will align with relevant EU rules relating to the placing on the market of manufactured goods [19].

From 1 January 2023, the UKCA marking must, in most cases, be affixed directly to the product [20]. You can still use the CE marking until then. However, the CE marking is only valid in Great Britain for areas where GB and EU rules remain the same. If the EU changes its rules and you CE mark your product on the basis of those new rules you will not be able to use the CE marking to sell in Great Britain, even before 31 December 2022 [20].

- **Effective countries.** Great Britain (England, Scotland, and Wales). Northern Ireland requires the CE mark [20].
- **Responsible authority.** Department for Business, Energy & Industrial Strategy
- **Statutory basis.** UK Statutory Instruments 2019/696 (Product Safety and Metrology etc. (Amendment etc.) (EU Exit) Regulations 2019)
- **Scope of UKCA marking.** Safety, health, environmental protection, electromagnetic compatibility and other regulations.
- **UKCA marking obligation for electronic products.** Mandatory.
- **Emissions tests required.** Yes.
- **Immunity tests required.** Yes.
- **EMC standards.** The British Standards Institution (BSI) maintains the EMC standards. Most EMC standards are identical to the EN standards of the EU. However, they use the prefix “BS” to indicate that they are standards adopted by the BSI as the UK’s national standards body [20].

J.14 United States of America

The Federal Communications Commission (FCC) mark (Fig. J.15) is the product conformity mark of the United States of America (USA).

- **Effective countries.** United States of America (USA).

Fig. J.15 The Federal Communications Commission (FCC) mark

- **Responsible authority.** Federal Communications Commission (FCC).
- **Statutory basis.** 47 CFR Part 15 (ITE) [31], 47 CFR Part 18 (ISM) [31].
- **Scope of FCC marking.** Electromagnetic Interference (EMI), Electromagnetic Compatibility (EMC).
- **FCC marking obligation for electronic products.** Applicable to electronic equipment.
- **Emissions tests required.** Yes.
- **Immunity tests required.** No (for FCC mark). However, in case of, e.g., military or medical equipment, the MIL or AAMI standards are applicable, which do require immunity tests.
- **EMC standards.** Emission limits: 47 CFR [31]. Test methods: FCC MP-5-1986 [17], IEEE/ANSI C63x.

  The Code of Federal Regulations (CFR), Title 47 (Telecommunications), Chapter I (Federal Communication Commission) contains the following parts, which can be seen as the United States (US) EMC regulations:

  – Part 15—Radio Frequency Devices
  – Part 18—Industrial, Scientific and Medical Equipment (ISM)
  – Part 68—Communication of Terminal Equipment of the Telephone Network
  – Part 87—Aviation Services
  – Part 90—Private Land Mobile Radio Services
  – Part 97—Amateur Radio Services

  One difference between the US regulations and the EU regulations is that the US regulations (FCC, Title 47, Chapter I) also specify the limits in the law, e.g., the conducted limits of unintentional radiators 47 CFR 15.107. Whereas in the EU directives, this is not the case and the test limits are specified in the EMC standards issued by the IEC and/or ISO organizations. On the other side, the testing methods for the US are mostly specified by the standards organization (e.g., ANSI, IEEE, or AAMI). Here are the most important EMC standards for the US:

  – **IEEE/ANSI C63.4** American National Standard for Methods of Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz. ANSI C63.4, conducted emission testing, conducting ground plane, digital equipment, electric field measurement, line impedance stabilization network, low-voltage electrical equipment, low-voltage electronic equipment, magnetic field measurement, normalized site attenuation, radiated emission testing, radio-noise emissions, radio-noise power, site attenuation, unintentional radiators.

- **IEEE/ANSI C63.10.** American National Standard of Procedures for Compliance Testing of Unlicensed Wireless Devices. The procedures for testing the compliance of a wide variety of unlicensed wireless transmitters (also called intentional radiators and license-exempt transmitters) including, but not limited to, remote control and security unlicensed wireless devices, frequency hopping and direct sequence spread spectrum devices, antipilferage devices, cordless telephones, medical unlicensed wireless devices, Unlicensed National Information Infrastructure (U-NII) devices, intrusion detectors, unlicensed wireless devices operating on frequencies below 30 MHz, automatic vehicle identification systems, and other unlicensed wireless devices authorized by a radio regulatory authority are covered in this standard. Excluded by this standard are test procedures for unlicensed wireless devices already covered in other published standards (e.g., Unlicensed Personal Communication Services (UPCS) devices).
- **IEEE/ANSI C63.15.** American National Standard Recommended Practice for the Immunity Measurement of Electrical and Electronic Equipment. This immunity testing and test instrumentation specifications recommended practice complements the procedures for making emission measurements as specified in ANSI C63.4 and in ANSI C63.10. These immunity test methods can be used by manufacturers who want to maximize product reliability and reduce customer complaints by improving the immunity of their products, beyond that required by applicable regulations, or by correcting problems experienced in deployment that are not related to regulatory requirements. This recommended practice generally covers the frequency range 30 Hz to 10 GHz.
- **IEEE/ANSI C63.17.** American National Standard Methods of Measurement of the Electromagnetic and Operational Compatibility of Unlicensed Personal Communications Services (UPCS) Devices. Specific test procedures for verifying the compliance of unlicensed personal communications services (UPCS) devices (including wide-band voice and data devices) are established including applicable regulatory requirements regarding radio-frequency emission levels and spectrum access procedures.
- **FCC MP-5-1986.** Methods of measurement of radio noise emissions from Industrial, Scientific and Medical (ISM) equipment).
- **ANSI/AAMI.** The US Food and Drug Administration (FDA) is responsible for defining the requirements for medical equipment and medical devices. The FDA refers to the Association for the Advancement of Medical Instrumentation (AAMI) Standards for compliance [30].
- **SAE**. SAE International (previously Society of Automotive Engineers (SAE)) is a US-based association and standards developing organization for engineering professionals in various industries, especially in the transportation industry. The SAE EMC standards are divided into two categories: SAE J1113-xx for components and SAE J551-xx for vehicles. Many SAE EMC standards were submitted to the international organizations ISO and CISPR for consideration as an international standard. Many SAE EMC standards have

become international standards and were therefore withdrawn. However, there are still active SAE EMC standards remaining, which were not (yet) adopted by ISO and CISPR.

– **MIL-STD.** MIL-STDs are US military standards. Here the two most important United States Department of Defense (DoD) EMC standards:

  MIL-STD-461. Requirements for the Control of Electromagnetic Interference Characteristics of Subsystems and Equipment. MIL-STD-461 contains emission and immunity requirements. Unlike FCC or IEEE/ANSI standards, MIL-STD-461 is not a legally binding EMC standard. MIL-STD-461 is a contractual standard, where test limits and emission levels can be negotiated and waivers are possible. This standard is best suited for items that have the following features: electronic enclosures that are no larger than an equipment rack, electrical interconnections that are discrete wiring harnesses between enclosures, and electrical power input derived from prime power sources. MIL-STD-461 should not be directly applied to items such as modules located inside electronic enclosures or entire platforms. The principles in MIL-STD-461 may be useful as a basis for developing suitable requirements for those applications.

  MIL-STD-464. Electromagnetic Environmental Effects (E3) Requirements for Systems. MIL-STD-464 is the DoD EMC standard which establishes electromagnetic environmental effects (E3) interface requirements and verification criteria for airborne, sea, space, and ground systems, including associated ordnance. MIL-STD-464 is applied at system or platform level.

– **RTCA DO-160** contains Environmental Conditions and Test Procedures for Airborne Equipment and is the minimum standard for the environmental testing of commercial avionics hardware. It is published by the Radio Technical Commission for Aeronautics (RTCA). Like the MIL-STD standards, RTCA DO-160 is a contractual standard (not legally binding). The European Organization for Civil Aviation Equipment (EUROCAE) works jointly with RTCA in the development of standards and publishes an identical standard called EUROCAE ED-14D [32].

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Appendix K
EMC Test Setups and Equipment

Testing leads to failure, and failure leads to understanding.

—Burt Rutan

This chapter presents typical EMC test setups and equipment. As an EMC design engineer, it is important to understand the EMC test equipment and setups. In particular:

- What are the frequency ranges, voltage levels, and field strength for emission measurements and immunity tests?
- What are the voltage levels, rise/fall-time durations, and energy of transient pulses (ESD, burst, surge)?
- What is the maximum permissible current at which harmonic order?
- What is the duration of an AC or DC power supply interruption or dip and how is a dip specified?

All this information gives guidance on how to design specific filters, shield barriers, or power supplies.

The author can accept no responsibility or liability for the content of this chapter and throughout this book. All information provided in this chapter and throughout the book are provided as is, with no warranties, and should not be used as a reference for compliance testing. Additionally, the test setups presented in this chapter and through this book are not legally binding. Please make sure to test your product according to the recognized EMC standards at an accredited EMC laboratory.

K.1 Pre-compliance vs. Compliance Testing

Before we jump into the topic of compliance EMC test setups and equipment, here is a summary and comparison between pre-compliance and compliance testing:

- **Pre-compliance testing and troubleshooting.** *Pre-compliance EMC testing* takes place at not accredited EMC test sites. Such pre-compliance testing is

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typically performed by the product development team itself with the goal to identify obvious weak spots of the product regarding EMC *before* the product will be tested at an accredited EMC test laboratory. Troubleshooting is often part of the EMC pre-compliance tests. How to create your own troubleshooting kit is wonderfully described in this book: [17].
Typical measurement and test equipment for pre-compliance tests and troubleshooting are:

– **Spectrum analyzers.** Measuring radiated RF emissions (near- and far-field) or conducted RF emissions (with Line Impedance Stabilization Network (LISN)). Be aware that spectrum analyzers are similar to EMI receivers (both measure signals in the frequency domain). However certified EMI receivers must be compliant with the requirements according to CISPR 16-1-1 [16] (dynamic range, quasi-peak-detector).
– **Oscilloscopes.** Oscilloscopes for identifying fast signal transients or crosstalk in the time-domain are often used for troubleshooting.
– **Uncertified test equipment.** Usually, uncertified test equipment is used for pre-compliance tests.

- **Compliance Testing.** *Compliance testing* takes place at accredited EMC test laboratories with calibrated and certified measurement and test equipment:

  – **EMI receivers.** *EMI receivers* are very similar to ordinary *spectrum analyzers* with some essential differences, e.g., regarding pre-selection filter, pre-amplifier, and quasi-peak detector (specified in CISPR 16-1-1).
  – **Fully or semi-anechoic chambers.** Radiated emission and immunity tests take place usually in fully or semi-anechoic chambers.
  – **Certified test equipment.** Certified test equipment is required for compliance testing. The specifications for certified test equipment (voltage and current generators, antennas, filters, test chambers) and their calibration is given in the respective EMC standards.

K.2 RF Emission: CISPR 11

The IEC/CISPR 11, EN 55011 (Industrial, scientific and medical (ISM) radio-frequency equipment—electromagnetic disturbance characteristics—limits and methods of measurement) is about: conducted and radiated emissions of signals in the frequency range of 9 kHz to 400 GHz.

K.2.1 CISPR 11: Applicability

Equipment covered by other CISPR product and product family emission standards are excluded from the scope of CISPR 11. CISPR 11 applies to industrial, scientific, and medical electrical equipment operating in the frequency range 0 Hz to 400 GHz and to domestic and similar appliances designed to generate and/or use locally radio-frequency energy. CISPR 11 standard covers emission requirements related to radio-frequency (RF) disturbances in the frequency range of 9 kHz to 400 GHz. Measurements need only be performed in frequency ranges where limits are specified.

CISPR 11 and CISPR 14: Induction cooking is the scope of CISPR 14-1 and CISPR 14-2. Microwave ovens are primarily the scope of CISPR 11 and CISPR 14-2 (however, be aware of multi-function equipment which may also be the scope of CISPR 14-1, e.g., for click measurement).

For ISM Radio-Frequency (RF) applications in the meaning of the definition found in the United Nations specialized agency for information and communication technologies (ITU) Radio Regulations [15], CISPR 11 covers emission requirements related to radio-frequency disturbances in the frequency range of 9 kHz to 18 GHz. Requirements for ISM RF lighting equipment and Ultra Violet (UV) radiators operating at frequencies within the *ISM frequency bands* as defined by the ITU Radio Regulations are contained in CISPR 11. To clarify things, here are some points about ISM bands:

- The industrial, scientific, and medical (ISM) radio bands are reserved internationally for the use of radio-frequency (RF) energy for industrial, scientific, and medical purposes other than telecommunications.
- ISM bands are defined by the ITU Radio Regulations (article 5) in footnotes 5.138, 5.150, and 5.280 of the Radio Regulations [15].
- Not all ISM frequency bands are harmonized worldwide. Some apply only for certain regions (see Figs. K.1 and K.2).

K.2.2 CISPR 11: Group 1, 2

There are two groups of ISM equipment defined in CISPR 11 [14]:

- **Group 1 (general purpose applications, class A or class B):** All equipment in the scope of CISPR 11 which is not classified as Group 2 equipment. Examples of Group 1 equipment:
  - Laboratory equipment
  - Medical electrical equipment
  - Scientific equipment
  - Semiconductor-converters

Centre frequency [MHz]	Frequency range [MHz]	Maximum radiation limit in CISPR 11	Number of appropriate footnote to the table of frequency allocation of the ITU Radio Regulations Resolution No. 63
6,780	6,765 – 6,795	Under consideration	5.138
13,560	13,553 – 13,567	Unrestricted	5.150
27,120	26,957 – 27,283	Unrestricted	5.150
40,680	40,66 – 40,70	Unrestricted	5.150
433,920	433,05 – 434,79	Under consideration	5.138 in Region 1, except contries in 5.280
915,000	902 – 928	Unrestricted	5.150 in Region 2 only
2 450	2 400 – 2 500	Unrestricted	5.150
5 800	5 725 – 5 875	Unrestricted	5.150
24 125	24 000 – 24 250	Unrestricted	5.150
61 250	61 000 – 61 500	Under consideration	5.138
122 500	122 000 – 123 000	Under consideration	5.138
245 000	244 000 – 246 000	Under consideration	5.138

Footnote 5.138 = Frequency bands are designated for ISM applications. The use of these frequency bands for ISM applications shall be subject to special authorization by the administration concerned, in agreement with other administrations whose radiocommunication services might be affected. In applying this provision, administrations shall have due regard to the latest relevant ITU-R Recommendations.

Footnote 5.150 = Frequency bands are designated for industrial, scientific and medical (ISM) applications. Radiocommunication services operating within these bands must accept harmful interference which may be caused by these applications. ISM equipment operating in these bands is subject to the provisions of No. 15.13.

Footnote 5.280 = In Germany, Austria, Bosnia and Herzegovina, Croatia, The Former Yugoslav Republic of Macedonia, Liechtenstein, Montenegro, Portugal, Serbia, Slovenia and Switzerland, the band 433.05-434.79 MHz (centre frequency 433.92 MHz) is designated for industrial, scientific and medical (ISM) applications. Radiocommunication services of these countries operating within this band must accept harmful interference which may be caused by these applications. ISM equipment operating in this band is subject to the provisions of No. 15.13.

Fig. K.1 ISM frequency bands by ITU [15]

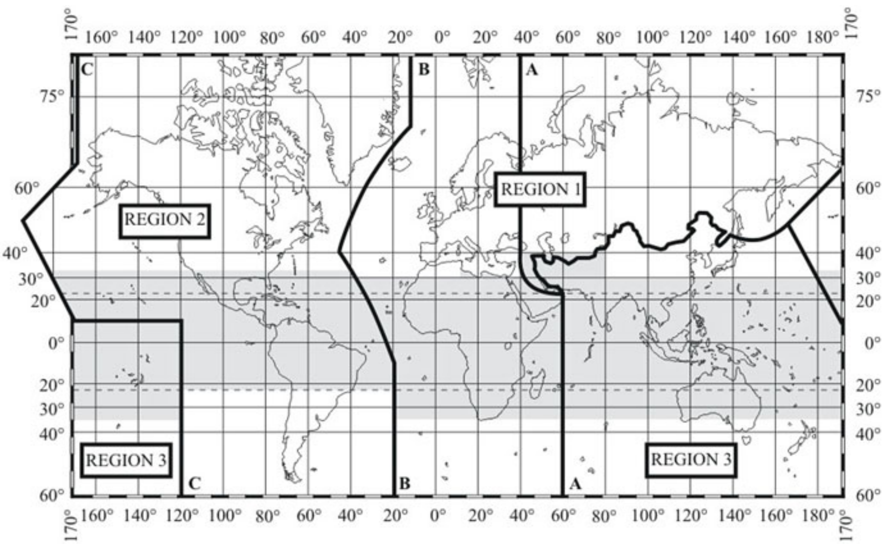


Fig. K.2 ISM frequency bands by ITU [2]

  - Industrial electric heating equipment with operating frequencies less than or equal to 9 kHz
  - Machine tools
  - Industrial process measurement and control equipment
  - Semiconductor manufacturing equipment
  - Switch mode power supplies

- **Group 2 (ISM RF applications, class A or class B):** All ISM RF equipment in which radio-frequency energy in the frequency range 9 kHz to 400 GHz is intentionally generated and used or only used locally, in the form of electromagnetic radiation, inductive and/or capacitive coupling, for the treatment of material, for inspection/analysis purposes, or for transfer of electromagnetic energy. Examples of Group 2 equipment:

  - Microwave-powered UV irradiating apparatus
  - Microwave lighting apparatus
  - Industrial induction heating equipment operating at frequencies above 9 kHz
  - Dielectric heating equipment Industrial microwave heating equipment
  - Arc welding equipment
  - Microwave ovens
  - Medical electrical equipment
  - Electric welding equipment
  - Electro-discharge machining (EDM) equipment
  - Demonstration models for education and training
  - Battery chargers and power supplies—wireless power transfer (WPT) mode

K.2.3 CISPR 11: Class A, B

There are two classes of ISM equipment defined in CISPR 11 [14] (Table K.1):

- **Class A** (higher emission limits, industrial): Class A devices are devices that are suitable for use in all areas other than residential and such areas, and they are connected to the public mains. Devices must have emissions below the limits of Class A, but the emissions may exceed the limits of Class B. For Class A equipment, the instructions for use accompanying the product shall contain the following text: *Caution: This equipment is not intended for use in residential environments and may not provide adequate protection to radio reception in such environments.*
- **Class B** (lower emission limits, residential): Class B devices are devices that are suitable for use in residential areas and such areas, and they are connected to the public mains.

Table K.1 CISPR 11 groups and classes

CISPR 11	Group 1 (All other ISM)	Group 2 (ISM with intentional radiated RF)
Class A (Industrial)	Nonintended RF emitters. Industrial environment.	Intentional RF emitters. Industrial environment.
Class B (Residential)	Nonintended RF emitters. Domestic environment.	Intentional RF emitters. Domestic environment.

K.2.4 CISPR 11: Test Setup

Test location [14]:

- **Class A.** Class A equipment may be measured either on a test site or in situ (at the installation site) as preferred by the manufacturer. Due to size, complexity, or operating conditions, some equipment may have to be measured in situ to show compliance with the radiation disturbance limits.
- **Class B.** Class B equipment shall be measured on a test site.

Measuring distance for radiated emissions [14]:

- **Class A.** On a test site, Class A equipment can be measured at a nominal distance d [m] of 3 m, 10 m or 30 m (Figs. K.3 and K.4). Class A Group 1 equipment can be measured in situ, where the measurement takes place at a distance of 30 m from the outer face of the exterior wall of the building in which the equipment is situated. Class A Group 2 equipment can be measured in situ, where the measurement distance d from the exterior wall of the building in which the equipment is situated equals $30\,\mathrm{m} + x/a$ [m] or 100 m whichever is smaller, provided that the measuring distance d [m] is within the boundary of the premises. In the case where the calculated distance d is beyond the boundary of the premises, the measuring distance d [m] equals x [m] or 30 m, whichever is longer. For the calculation of the above values:
  - x [m] is the nearest distance between the exterior wall of the building in which the equipment is situated and the boundary of the user's premises in each measuring direction.
  - $a = 2.5$ for frequencies lower than 1 MHz.
  - $a = 4.5$ for frequencies equal to or higher than 1 MHz.
- **Class B.** On a test site, Class B equipment can be measured at a nominal distance d [m] of 3 m or 10 m.
- **d** $<$ **10 m.** In the frequency range 30 MHz to 1000 MHz, a distance less than 10 m is allowed only for equipment which complies with the definition for small size equipment. Small size equipment is either positioned on a table top or standing on

the floor which, including its cables, fits in an imaginary cylindrical test volume of 1.2 m in diameter and 1.5 m height (to ground plane).

Consistent with typical applications of the equipment under test, the level of the disturbance shall be maximized by varying the configuration of the equipment.

K.2.5 CISPR 11: Test Limits

Be aware that CISPR 11 limits do not apply for ISM frequency bands (defined by ITU). Outside of ITU-designated ISM bands, the limits for the disturbance voltage and radiation disturbance in CISPR 11 apply.

K.2.5.1 CISPR 11: Test Limits at Test Site

There are limits for conducted and radiated emissions specified in CISPR 11. Here are the limits for measurements performed on a test site (not in situ).
CISPR 11 Group 1 [14]:

- 9 kHz to 150 kHz. No limits specified.
- 150 kHz to 30 MHz. Conducted emissions limits—public mains power ports—Group 1 equipment (Class A and B). Conducted emissions limits—DC power ports—Group 1 equipment (Class A and B) (Figs. K.5, K.6 and K.7).
- 30 MHz to 1 GHz. Radiated emissions limits—Group 1 equipment (Class A and B) (Figs. K.8 and K.9).
- 1 GHz to 18 GHz. No limits specified.
- 18 GHz to 400 GHz. No limits specified.

CISPR 11 Group 2 [14]:

- 9 kHz to 150 kHz. No limits specified.
- 150 kHz to 30 MHz. Conducted emissions limits—public mains power ports—Group 2 equipment (Class A and B) (Figs. K.5, K.6 and K.7).
- 30 MHz to 1 GHz. Radiated emissions limits—Group 2 equipment (Class A and B) (Figs. K.8 and K.9).
- 1 GHz to 18 GHz. The limits in the frequency range 1 GHz to 18 GHz apply only to Group 2 equipment operating at frequencies above 400 MHz. The evaluation of the radiated emission limits is complicated and would go beyond the constraints of this website.
- 18 GHz to 400 GHz. No limits specified.

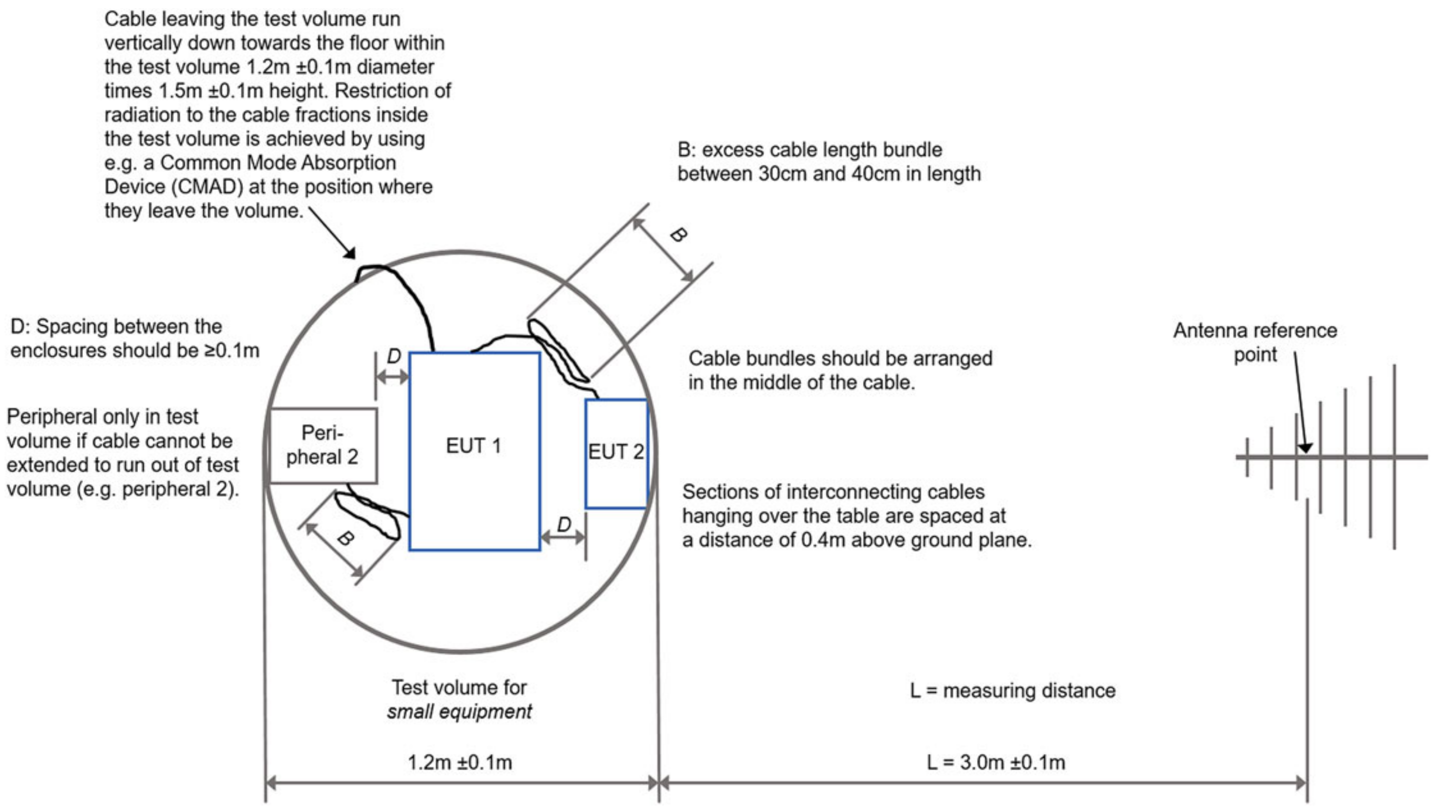


Fig. K.3 Example of a typical setup for measurements of radiated disturbances from a tabletop EUT. The measurement arrangement shall be typical of normal installation practice and centered to the turntable's vertical axis [14]

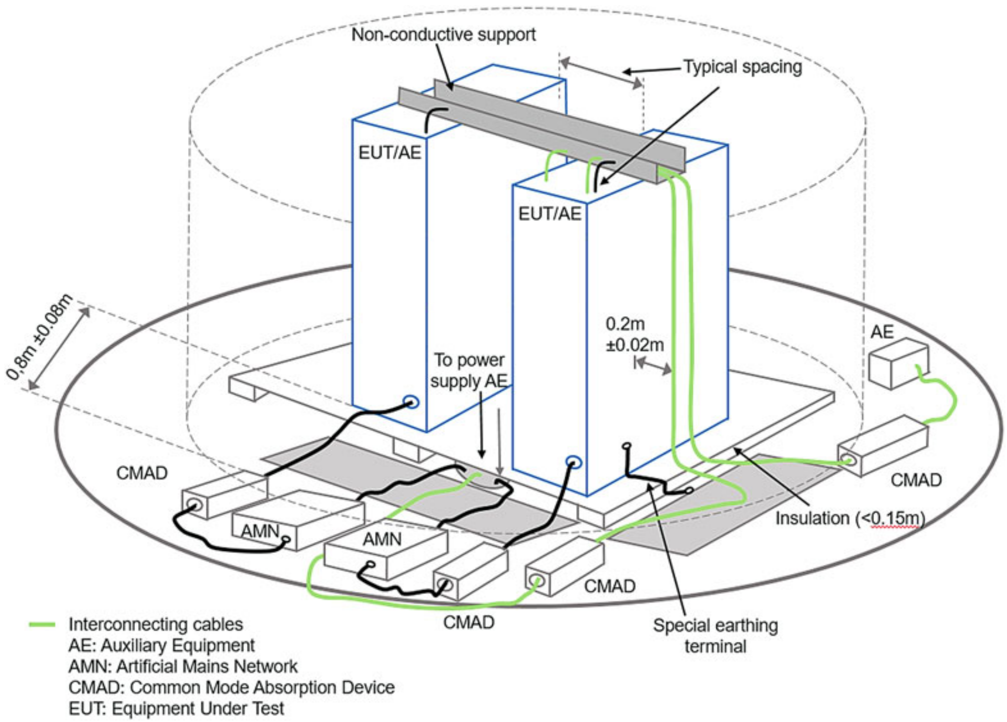


Fig. K.4 Example of a typical unified test set up for floor standing equipment suitable for measurement of conducted as well as radiated disturbances. Further examples of typical arrangements of the EUT and associated peripherals are given in CISPR 16-2-3 and CISPR 16-2-1 [14]

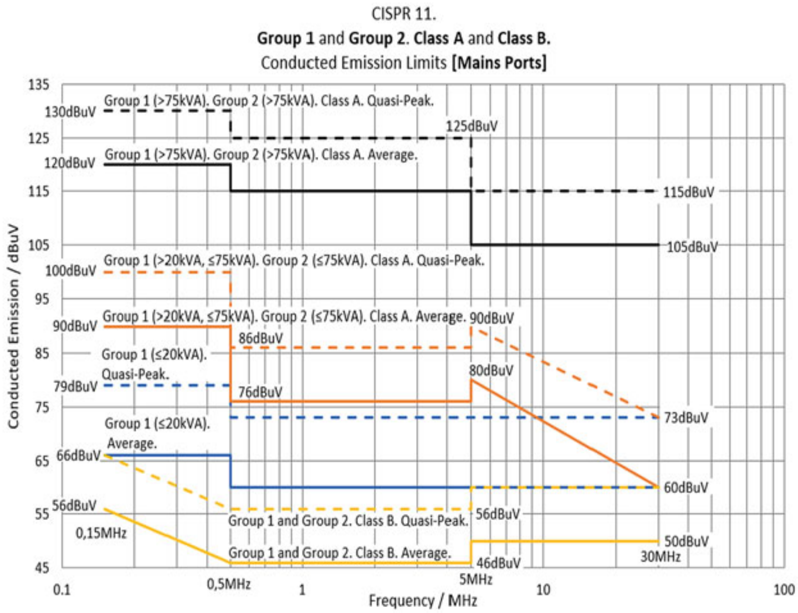


Fig. K.5 CISPR 11 CE limits for mains ports at test site [14]

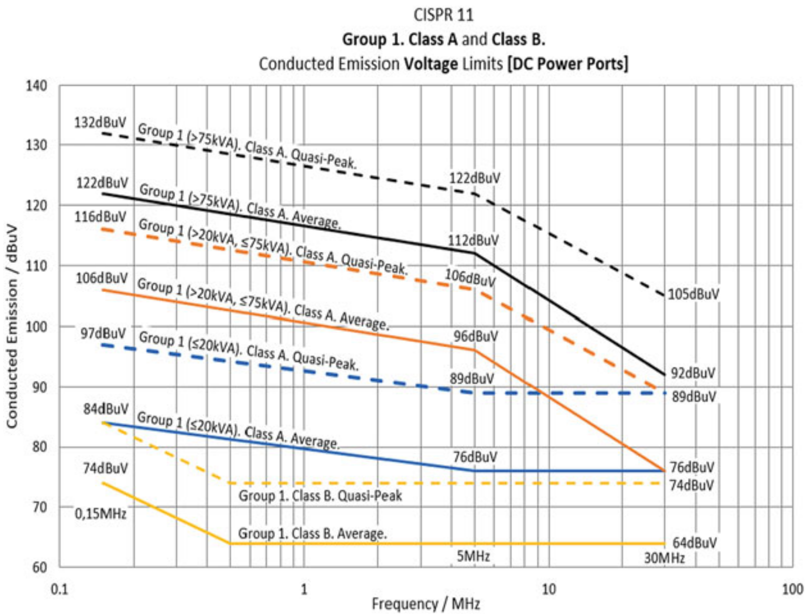


Fig. K.6 CISPR 11 CE voltage limits for DC ports for group 1 at test site [14]

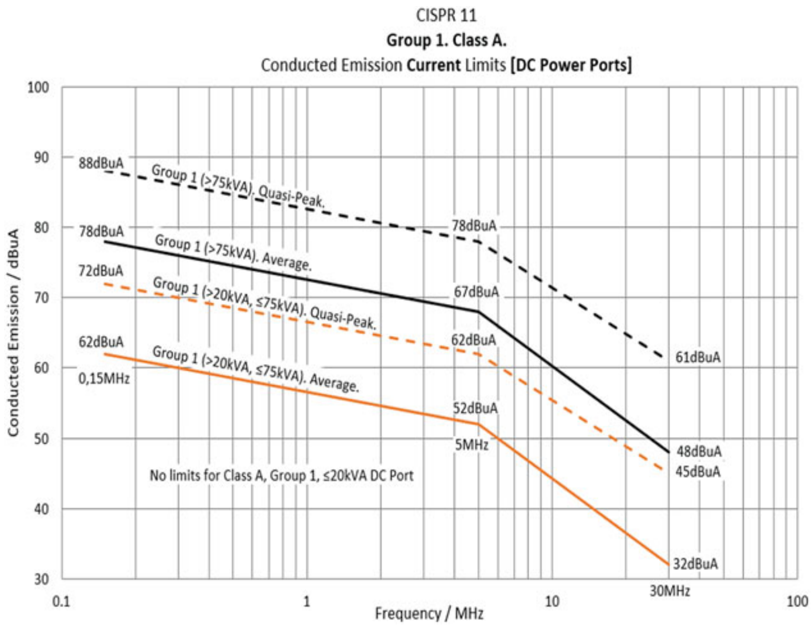


Fig. K.7 CISPR 11 CE current limits for DC ports for group 1 at test site [14]

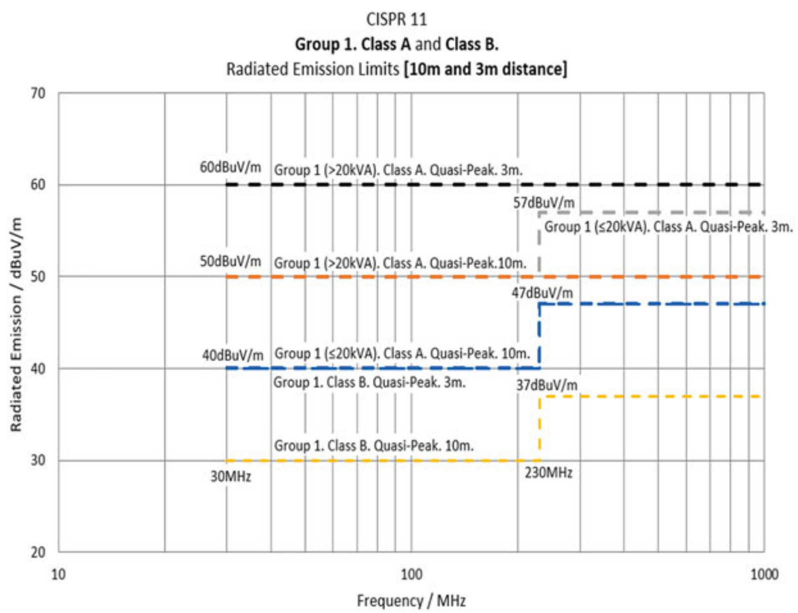


Fig. K.8 CISPR 11 RE limits for group 1 at test site [14]

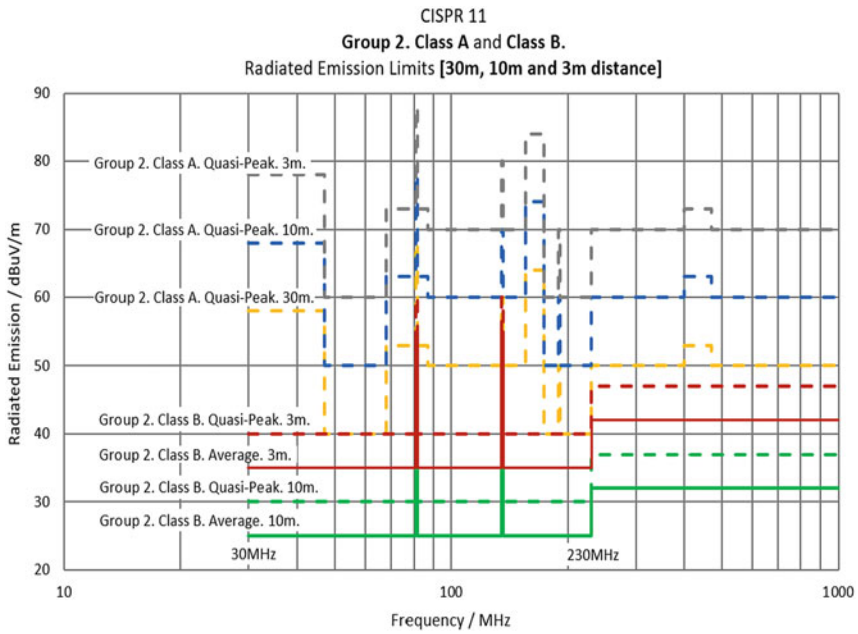


Fig. K.9 CISPR 11 RE limits for group 2 at test site [14]

K.2.5.2 CISPR 11: Test Limits In Situ

Under in situ conditions, an assessment of conducted disturbances is not required.
CISPR 11 Group 1 [14]:

- 150 kHz to 30 MHz. Magnetic field radiation limits—Group 1 Class A (Fig. K.10).
- 30 MHz to 1000 MHz. Electric field radiation limits—for Group 1 Class A (Fig. K.11).

CISPR 11 Group 2 [14]:

- 150 kHz to 30 MHz. Magnetic field radiation—limits for Group 2 Class A (Fig. K.12).
- 30 MHz to 1000 MHz. Electric field radiation—limits for Group 2 Class A (Fig. K.13).

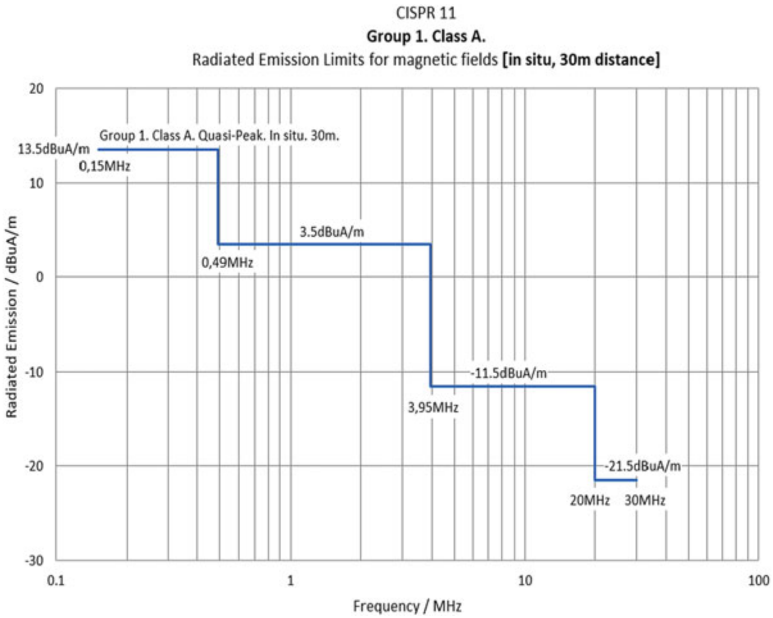


Fig. K.10 CISPR 11 magnetic field RE limits for group 1 in situ [14]

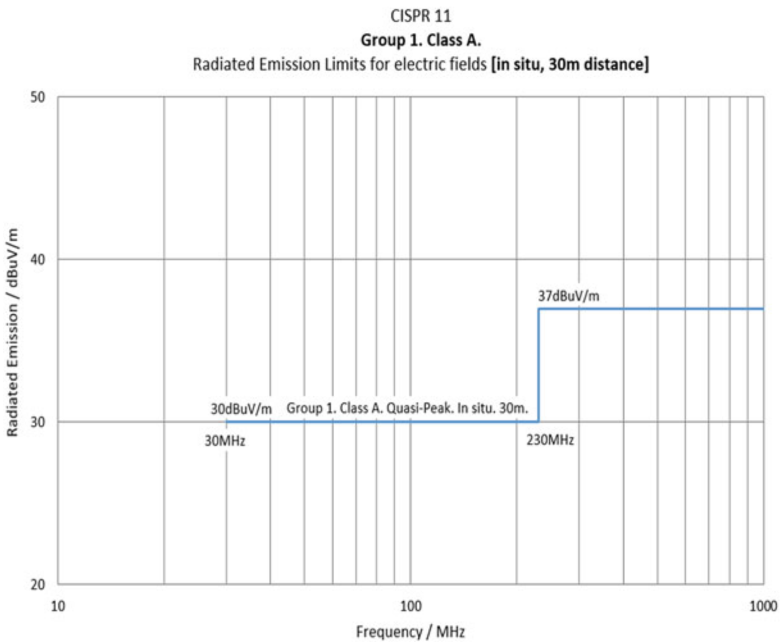


Fig. K.11 CISPR 11 electric field RE limits for group 1 in situ [14]

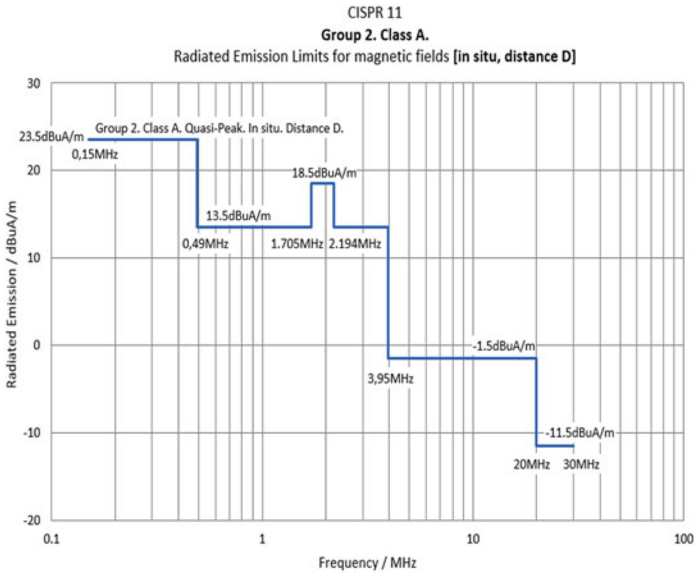


Fig. K.12 CISPR 11 magnetic field RE limits for group 2 in situ [14]

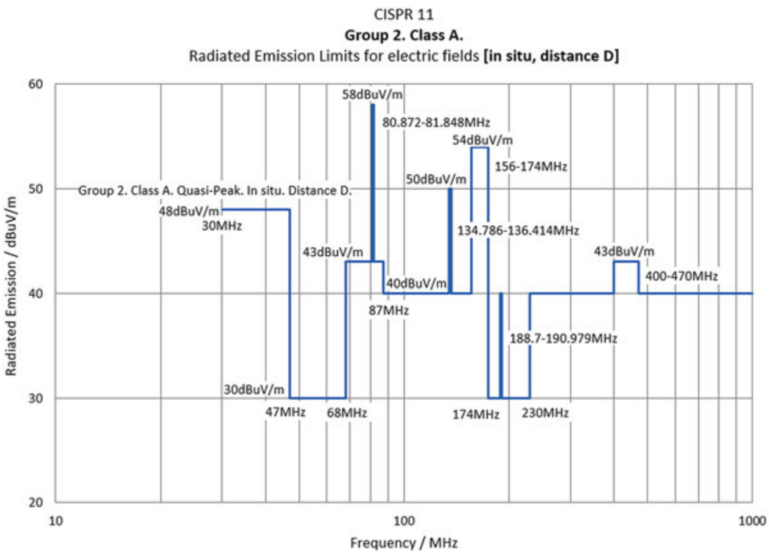


Fig. K.13 CISPR 11 electric field RE limits for group 2 in situ [14]

K.3 RF Emission: CISPR 32

The IEC/CISPR 32, EN 55032 (Electromagnetic compatibility of multimedia equipment—Emission requirements) is about: conducted and radiated emissions of signals in the frequency range of 9 kHz to 400 GHz. CISPR 32 replaced CISPR 13 and the popular CISPR 22 in March 2017.

K.3.1 CISPR 32: Applicability

CISPR 32 applies to multimedia equipment (MME) and having a rated RMS AC or DC supply voltage not exceeding 600 V. Equipment within the scope of CISPR 13 or CISPR 22 is within the scope of CISPR 32. MME intended primarily for professional use is within the scope of CISPR 32. The radiated emission requirements in CISPR 32 are not intended to be applicable to the intentional transmissions from a radio transmitter as defined by the ITU, nor any spurious emissions related to these intentional transmissions. Equipment for which emission requirements in the frequency range covered by CISPR 32 are explicitly formulated in other CISPR publications (except CISPR 13 and CISPR 22) are excluded from the scope of this publication. CISPR 32 does not contain requirements for in situ assessment (in other words, the tests have to be done in an EMC test laboratory). The objectives of CISPR 32 publication are [3]:

- To establish requirements which provide an adequate level of protection of the radio spectrum, allowing radio services to operate as intended in the frequency range 9 kHz to 400 GHz.
- To specify procedures to ensure the reproducibility of measurement and the repeatability of results.

CISPR 32 is sometimes referenced by other product and product family standards, outside of the scope defined above.

K.3.2 CISPR 32: Class A, B

There are two classes of Information Technology Equipment (ITE) defined in CISPR 32 [3]:

- **Class A** (higher emission limits, industrial): Devices must have emissions that are below the limits of Class A, but the emissions exceed the limits of Class B. Class A devices shall have a warning notice in their manual (e.g., “Warning! This is a Class A device. This device may cause radio interference in residential areas; in this case, the operator may be required to take appropriate measures.”).

- **Class B** (lower emission limits, commercial): Devices must have emissions that are below the limits of Class B. This is applicable for devices that are used in a residual and domestic environment. In other words, commercial devices. For example:
  - No permanent location (e.g., battery-powered devices)
  - Telecommunication terminal equipment
  - Personal computers

K.3.3 CISPR 32: Limits

There are limits for conducted and radiated emissions specified in CISPR 32. They can be found in the pictures below. The limits for quasi-peak and average must not be exceeded by Class A devices (blue) or Class B devices (orange), respectively.

There are also the radiated emission limits for 3 m measurement distance added, which are in general $20 \cdot \log_{10}(10\,\text{m}/3\,\text{m}) = 10.5\,\text{dB}$ higher compared to the specified limits for 10 m distance. If the measurement takes place at 1 m distance (1 m between Equipment Under Test (EUT) and antenna), the limits are $20 \cdot \log_{10}(10\,\text{m}/1\,\text{m}) = 20\,\text{dB}$ higher than the 10 m limits.

- Conducted emission limits [mains ports] (Fig. K.14)
- Conducted emission voltage limits [telecom/LAN ports] (Fig. K.15)
- Conducted emission current limits [telecom/LAN ports] (Fig. K.16)
- Radiated emission limits 30−1000 MHz [10 m vs. 3 m] (Fig. K.17)
- Radiated emission limits above 1 GHz (Fig. K.18)

K.4 Harmonic Currents: IEC 61000-3-2

The IEC 61000-3-2 (Limits for harmonic current emissions—equipment input current $\leq$ 16 A per phase) is a basic EMC publication which deals with the limitation of harmonic currents injected into the public mains supply system. It specifies limits of harmonic components of the input current which can be produced by equipment tested under specified conditions. The objective of IEC 61000-3-2 is to set limits for harmonic emissions of equipment within its scope so that, with due allowance for the emissions from other equipment, compliance with the limits ensures that harmonic disturbance levels do not exceed the compatibility levels defined in IEC 61000-2-2.

Professional equipment that does not comply with the requirements of IEC 61000-3-2 can be permitted to be connected to certain types of low voltage supplies if the instruction manual contains a requirement to ask the supply utility for

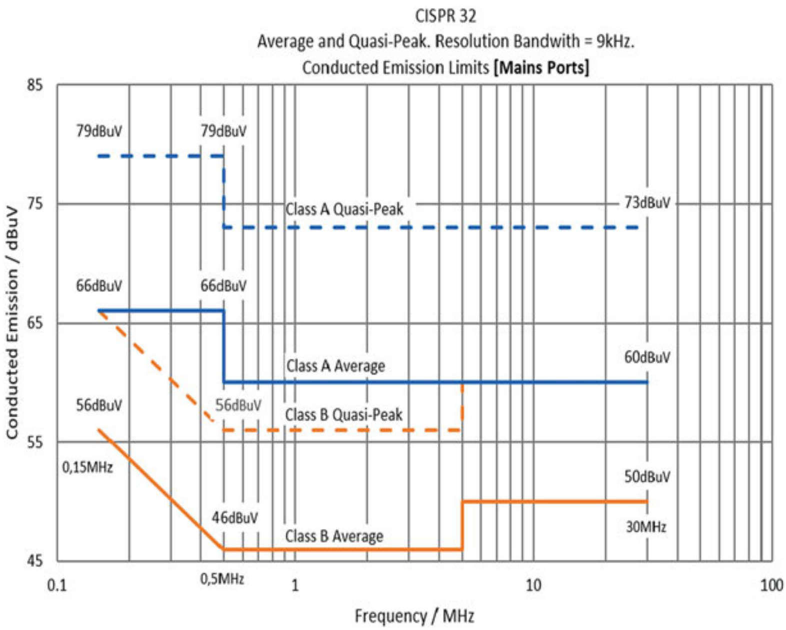


Fig. K.14 CISPR 32 CE limits for mains ports [3]

permission to connect. Recommendations concerning this aspect are contained in IEC 61000-3-12.

Figure K.19 shows an example current curve of a lightning equipment with dimming control. The current curve shows a phase shift and harmonic distortions.

K.4.1 IEC 61000-3-2: Applicability

IEC 61000-3-2 applies to:

- Apparatus intended to be connected to public low-voltage distribution systems (50 Hz or 60 Hz, 220/380 V, 230/400 V, and 240/415 V). For systems with nominal voltages less than but not equal to 220 V (line-to-neutral), the limits have not yet been considered.
- Apparatus having a rated input current up to and including 16 A per phase.
- Arc welding equipment which is not professional equipment, with a rated input current up to and including 16 A per phase. Arc welding equipment intended for professional use, as specified in IEC 60974-1, is excluded from IEC 61000-3-2 and can be subject to installation restrictions as indicated in IEC 61000-3-12.

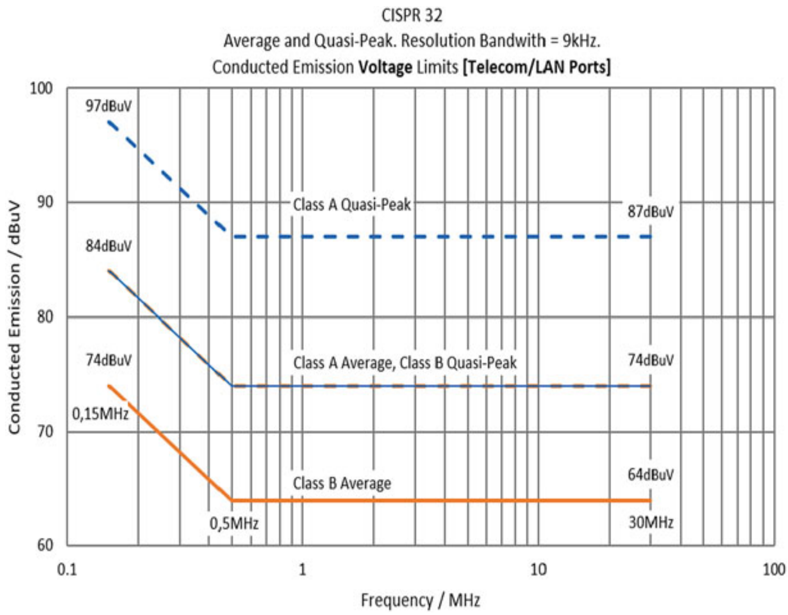


Fig. K.15 CISPR 32 CE voltage limits for telecom/LAN ports [3]

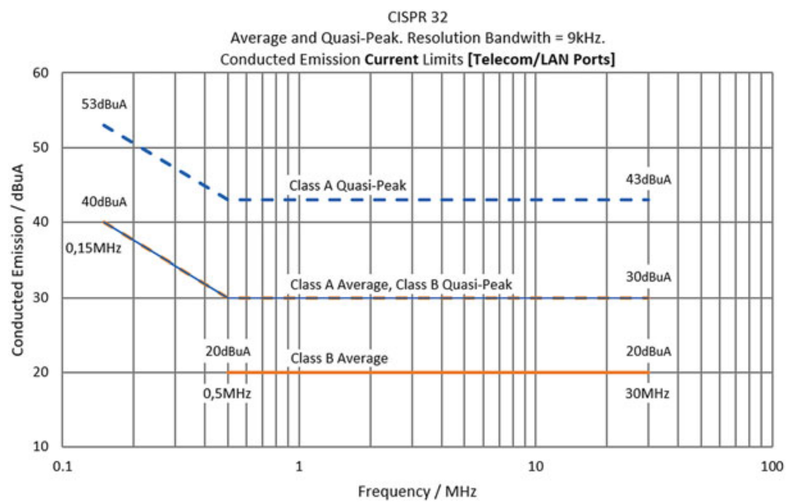


Fig. K.16 CISPR 32 CE current limits for telecom/LAN ports [3]

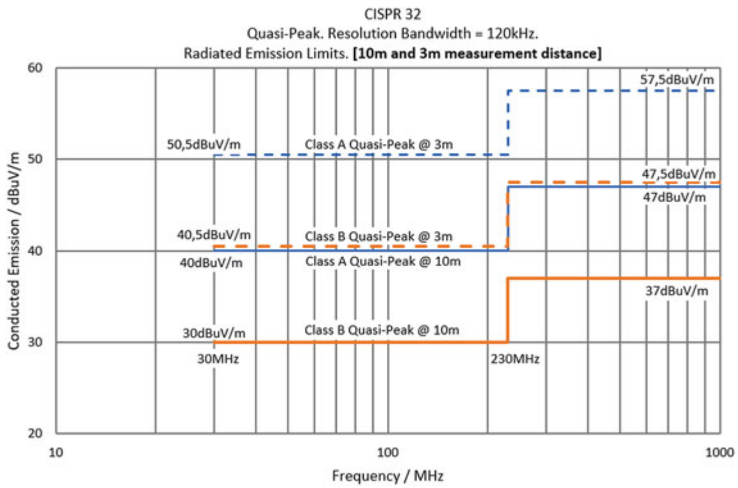


Fig. K.17 CISPR 32 RE limits up to 1 GHz [3]

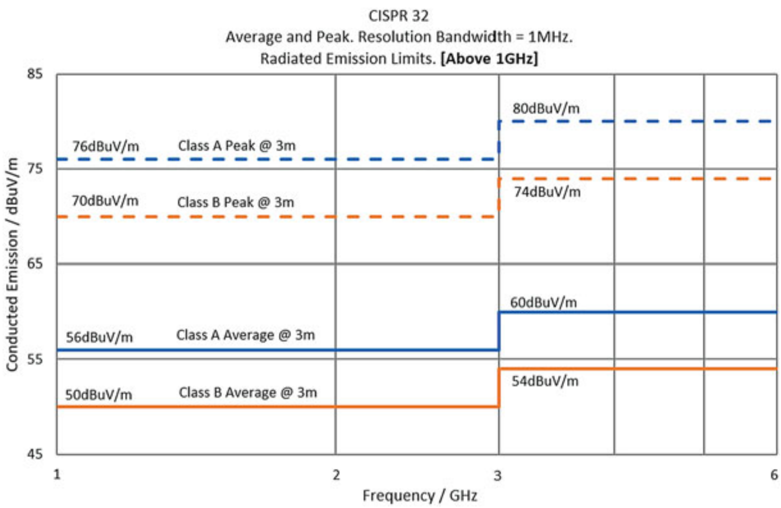


Fig. K.18 CISPR 32 RE limits above 1 GHz [3]

K.4.2 IEC 61000-3-2: Class A, B, C, D

There are four classes of equipment defined [5]:

- **Class A:** Equipment not specified as belonging to Class B, C, or D shall be considered as Class A equipment. Some examples of Class A equipment:
  - Balanced three-phase equipment.

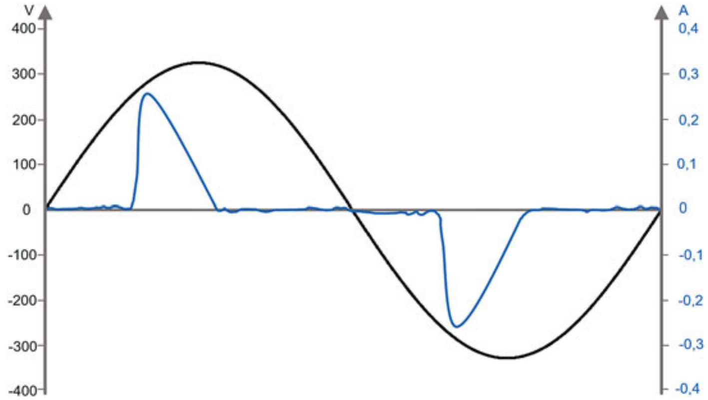


Fig. K.19 Public mains supply current distortion example [5]

- Household appliances, excluding those specified as Class B, C, or D.
- Vacuum cleaners.
- High pressure cleaners.
- Tools, excluding portable tools.
- Independent phase control dimmers.
- Audio equipment.
- Professional luminaires for stage lighting and studios.

- **Class B:** Portable tools and arc welding equipment which is not professional equipment.
- **Class C:** Lighting equipment.
- **Class D:** Equipment having a specified power consumption less than or equal to 600 W, of the following types:
  - Personal computers and personal computer monitors.
  - Television receivers.
  - Refrigerators and freezers having one or more variable-speed drives to control compressor motor(s).

K.4.3 IEC 61000-3-2: Test Setup

Figure K.20 shows an example of a single-phase measurement circuit. The measurement equipment shall be compliant with IEC 61000-4-7:

- S: Power supply source
- G: Open-loop voltage of the supply source
- M: Measurement equipment
- EUT: Equipment under test

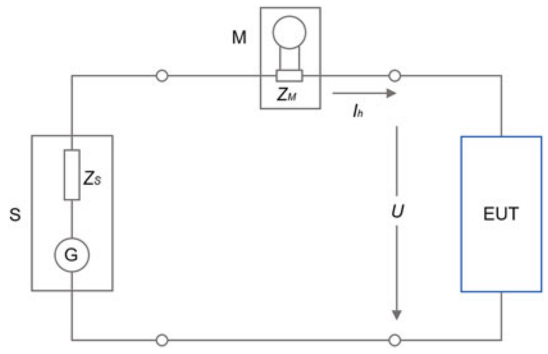


Fig. K.20 Harmonic currents test setup

- U: Test voltage
- I_h: Harmonic component of order h of the line current
- Z_M: Input impedance of measurement equipment
- Z_S: Internal impedance of the supply source

K.4.4 IEC 61000-3-2: Limits

The following categories of equipment are no limits specified in IEC 61000-3-2:

- Lighting equipment with a rated power < 5 W.
- Equipment with a rated power of ≤ 75 W, other than lighting equipment.
- Professional equipment with a total rated power > 1 kW.
- Symmetrically controlled heating elements with a rated ≤ 200 W.
- Certain types of independent phase control dimmers.

Before we have a look at the limits, have a look at Eqs. K.1 and K.2. According to IEC 61000-3-2, the total harmonic current (THC) is equal to the total RMS value of the harmonic current components of orders 2 to 40:

$$\text{THC} = \sqrt{\sum_{h=2}^{40} I_h^2} \tag{K.1}$$

According to IEC 61000-3-2, the total harmonic distortion (THD) is defined as the ratio of the RMS value of the sum of the harmonic components (in this context, harmonic current components I_h [A] of orders 2 to 40) to the RMS value of the fundamental component I_1 [A], expressed as:

$$\text{THD} = \sqrt{\sum_{h=2}^{40} \left(\frac{I_h}{I_1}\right)^2} \tag{K.2}$$

Table K.2 IEC 61000-3-2 class A limits [5]

IEC 61000-3-2. Limits for Class A equipment	
Harmonic order *h*	**Maximum permissible harmonic current** *A*
Odd harmonics	
3	2.30
5	1.14
7	0.77
9	0.40
11	0.33
13	0.21
$15 \le h \le 39$	$0.15 \cdot \frac{15}{h}$
Even harmonics	
2	1.08
4	0.43
6	0.30
$8 \le h \le 40$	$0.23 \cdot \frac{8}{h}$

IEC 61000-3-2 defines different limits depending on the equipment class:

- **Class A.** Limits for Class A equipment is shown in the Table K.2.
- **Class B.** Limits for Class B equipment are Class A limits multiplied by factor 1.5.
- **Class C.**
  - Rated power $\geq$ 5 W and $\leq$ 25 W.

    Harmonic currents shall not exceed the power-related limits of Class D (column 2 in Table K.4).
    The third harmonic current, expressed as a percentage of the fundamental current, shall not exceed 86% and the fifth harmonic current shall not exceed 61%. In addition, the waveform of the input current shall be such that it reaches the 5% current threshold before or at 60°, has its peak value before or at 65°, and does not fall below the 5% current threshold before 90°, referenced to any zero crossing of the fundamental supply voltage. The current threshold is 5% of the highest absolute peak value that occurs in the measurement window, and the phase angle measurements are made on the cycle that includes this absolute peak value. Components of current with frequencies above 9 kHz shall not influence this evaluation (a filter similar to the one described in IEC 61000-4-7:2002 may be used).
    The THD (formula can be found above) shall not exceed 70%. The third-order harmonic current, expressed as a percentage of the fundamental current, shall not exceed 35%, the fifth-order current shall not exceed 25%, the seventh-order current shall not exceed 30%, the ninth- and eleventh-order currents shall not exceed 20%, and the second-order current shall not exceed 5%.

  - Rated power >25 W.

Luminaires. For luminaires with incandescent lamps and built-in phase control dimming having a rated power greater than 25 W, the harmonics of the input current shall not exceed the limits of Class A equipment.
Any other lightning equipment. For any other lighting equipment having a rated power greater than 25 W, the harmonics of the input current shall not exceed the relative limits specified for Class C in Table K.3.

- **Class D.** Limits for Class D equipment are shown in Table K.4.

Table K.3 IEC 61000-3-2 class C limits [5]

IEC 61000-3-2. Limits for Class C equipment [a]	
Harmonic order h	**Maximum permissible harmonic current expressed as a percentage of the input current at the fundamental frequency** %
2	2
3	$30 \cdot \lambda^{b}$
5	10
7	7
9	5
$11 \le h \le 39$ (odd harmonics only)	3

[a] For some Class C products, other emission limits apply
[b] λ is the ciruict power factor

Table K.4 IEC 61000-3-2 class D limits [5]

IEC 61000-3-2. Limits for Class D equipment.		
Harmonic order h	**Maximum permissible harmonic current per watt** mA/W	**Maximum permissible harmonic current** A
3	3.40	2.30
5	1.90	1.14
7	1.00	0.77
9	0.5	0.40
11	0.35	0.33
$13 \le h \le 39$ (odd harmonics only)	$\frac{3.85}{h}$	$0.15 \cdot \frac{15}{h}$

K.5 Flicker: IEC 61000-3-3

IEC 61000-3-3 (limitation of voltage changes, voltage fluctuations, and flicker in public low-voltage supply systems, for equipment with rated current $\leq$ 16 A per phase and not subject to conditional connection) is a Product Family Standard about the limitation of voltage fluctuations and flicker impressed on the public low-voltage system. It specifies limits of voltage changes that may be produced by equipment tested under specified conditions and gives guidance on assessment methods.

The definition of flicker by IEC is as follows: Impression of unsteadiness of visual sensation induced by a light stimulus whose luminance or spectral distribution fluctuates with time [1].

Flicker can be considered a symptom resulting from the modulation of a load and its effect on its terminal voltage. While connected to a voltage source with finite impedance, any load modulation will cause voltage fluctuations on the supply line.

K.5.1 IEC 61000-3-3: Applicability

IEC 61000-3-3 applies to electrical and electronic equipment having an input current equal to or less than 16 A per phase, intended to be connected to public low-voltage distribution systems of between 220 V and 250 V line to neutral at 50 Hz, and not subject to conditional connection [6].

- Should a device fail the limits specified in IEC 61000-3-3 tested with the specified source impedance $\underline{Z}_{ref}$ [Ω], it may be retested to show conformity with IEC 61000-3-11. Part 3-11 is applicable to all equipment with rated input $\leq$ 75 A per phase and is subject to conditional connections.
- Devices that are unlikely to produce significant voltage fluctuations or flicker do not need to be tested. However, it may be necessary to clarify whether significant voltage fluctuations are generated with a certain probability by evaluating the circuit diagram and the device's specification and by a short functional test.
- For power grids with a voltage of less than 220 V, and/or with a frequency of 60 Hz, the limits and values for reference impedances are still under consideration.

$\underline{Z}_{ref}$ [Ω] is the internationally agreed reference source impedance for low-voltage supply networks.

K.5.2 IEC 61000-3-3: Test Setup

The test setup consists of [6] (Fig. K.21):

- G: Voltage source
- EUT: Equipment under test
- M: Measuring equipment
- S: Supply source including reference impedance $\underline{Z}_{ref}$ [Ω] and voltage generator output impedance:
  - $R_A = 0.24\,\Omega$, $jX_A = 0.15\,\Omega$ at 50 Hz
  - $R_N = 0.16\,\Omega$, $jX_N = 0.10\,\Omega$ at 50 Hz

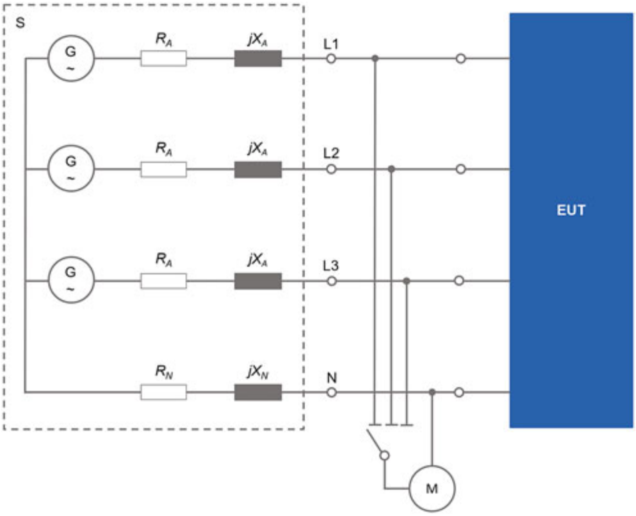


Fig. K.21 Flicker test setup [6]

K.5.3 IEC 61000-3-3: Limits

First, flicker limits are mainly based on the subjective perception of flicker in the light of 230 V/60 W incandescent lamp, which is caused by fluctuations in the supply voltage. Second, there are some terms to explain [6]:

- P_{st} is the calculated short-term flicker as per IEC 61000-4-15, unless otherwise specified the P_{st} evaluation period is 10 min.
- P_{lt} is the calculated long-term flicker as per IEC 61000-4-15. Unless otherwise specified, the P_{lt} evaluation period is 2 h ($N = 12$ in the formula below).

Commonly known as d values, voltage change characteristics constitute three separate parameters—$d(t)$, d_c, d_{max} (Figs. K.22 and K.23):

- $\mathbf{d(t)}$: Time function of the relative root mean square (RMS) half period voltage change for each half period between zero-crossings of the voltage source, expected during steady-state voltage conditions.
- $\mathbf{d_c}$: Maximum steady-state voltage change during an observation period
- $\mathbf{d_{max}}$: Maximum absolute voltage change during observation period
- $\mathbf{T_{max}}$: Maximum time that the half period RMS voltage exceeds the limit for DC. During a voltage change event, the T_{max} [sec] value is accumulated until a new steady-state condition is established.

The IEC 61000-3-3 limits are as follows:

- $\mathbf{P_{st}}$: Short-term flicker value P_{st} must be less than or equal to 1.0.
- $\mathbf{P_{lt}}$: Long-term flicker value P_{lt} must be less than or equal to 0.65.

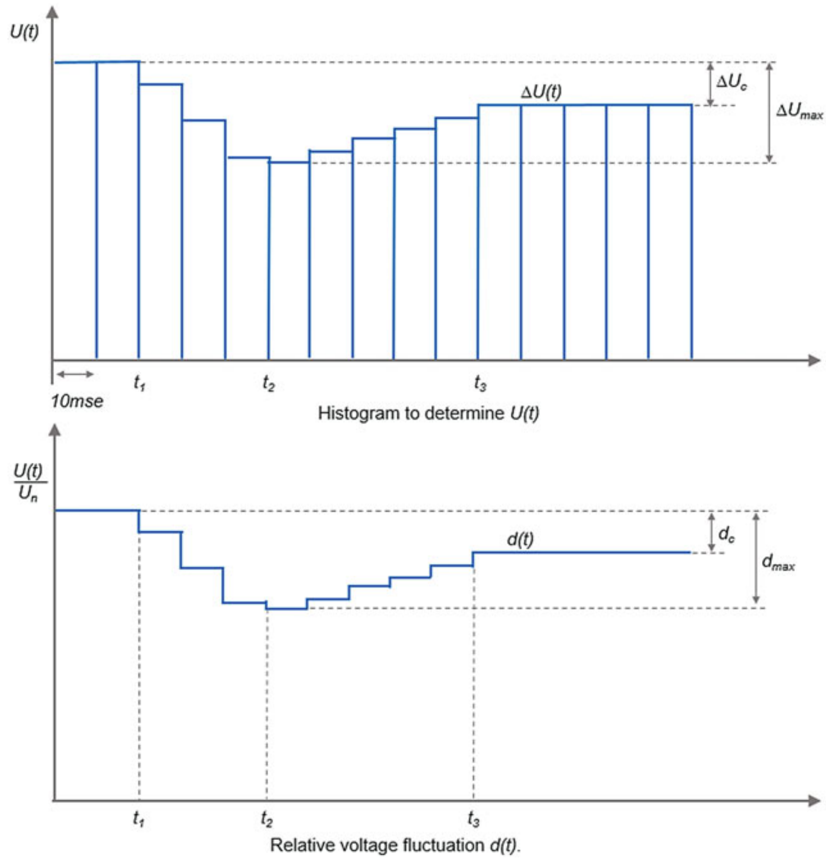


Fig. K.22 Flicker voltage fluctuation terms [6]

- $\mathbf{d(t), T_{max}}$: Accumulated time of $d(t)$ with a deviation exceeding 3.3% during a single voltage change at the equipment under test terminals must not exceed 500 msec.
- $\mathbf{d_c}$: The maximum relative steady-state voltage change DC must not exceed 3.3%.
- $\mathbf{d_{max}}$: Maximum relative voltage change (between two half periods) shall not exceed:
  - 4% without additional conditions
  - 6% for equipment that is:

    switched manually

    switched automatically more than twice a day and is also fitted with a delayed restart not less than a few tens of seconds. Alternatively a manual restart after a power supply interruption.

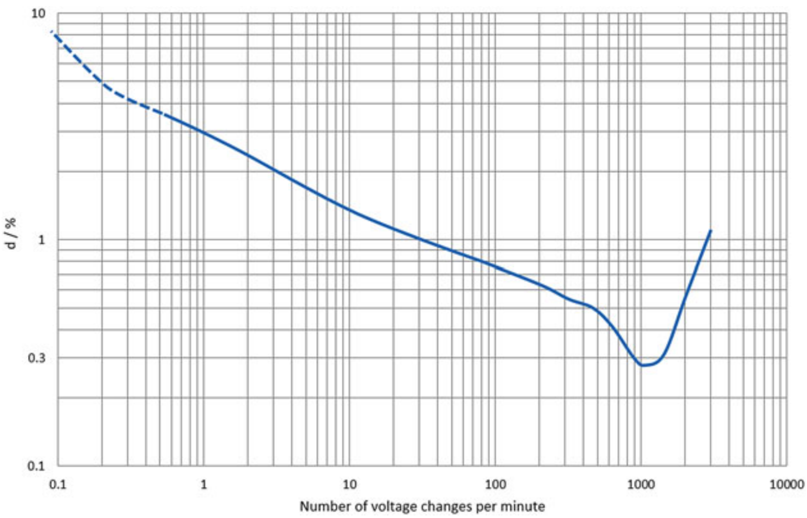


Fig. K.23 Flicker $P_{st} = 1$−curve for rectangular and equidistant voltage changes [6]. The concept behind this curve: If the performance of the EUT lies under the $P_{st} = 1$−curve, then the disturbance to the supply network is deemed to be acceptable in the short term. Remark: 1200 voltage fluctuations per minute lead to a 10 Hz flicker

– 7% for equipment that is:

  attended while in use

  switched on automatically or intended to be switched on automatically no more than twice per day. Must also be fitted with a delayed restart of not less than a few tens of seconds (or manual restart) after a power supply interruption.

K.6 ESD: IEC 61000-4-2

The IEC/EN 61000-4-2 EMC standard is about immunity testing (measurement techniques) against electrostatic discharge (ESD). This standard is part of the Basic EMC Publications.

IEC 61000-4-2 specifies immunity to *electrostatic discharge* from operating personnel and neighboring objects. The standard defines requirements, test methods, and test levels. The purpose of IEC 61000-4-2 is to establish a general and reproducible basis for determining the performance of electrical and/or electronic equipment when exposed to ESD. It also includes discharges of static electricity from persons to objects near the equipment under test.

K.6.1 IEC 61000-4-2: Test Setup

There are two test methods defined, whereas the first one is the most commonly applied because post-installation ESD tests may significantly decrease the mean time between failure (MTTF) of the EUT:

- Type conformity test performed in the laboratories.
- Test at the place of installation of the EUT.

The test setup described in the following refers to type conformity test performed in the laboratory. The following environmental conditions must be met during ESD testing [8]:

- Temperature: 15 °C to 35 °C
- Relative humidity: 30% to 60%
- Atmospheric pressure: 86 kPa (860 mBar) to 106 kPa (1060 mBar)

The test setup comprises the ESD test generator, the equipment under test (EUT), and its auxiliary equipment (AE) necessary to perform direct and indirect application of discharges to the EUT as applicable, in the following manner [8]:

- **Contact discharge** to conductive surfaces of the EUT and to neighboring coupling planes of the EUT (horizontal coupling plane HCP, vertical coupling plane VCP).
- **Air discharge** to insulating surfaces of the EUT.

Typical test setups for tabletop devices and floor-mounted apparatuses (with earth connection) are shown in Figs. K.24 and K.25.

The ESD generator emulates an electrical discharge of the human body. The storage capacity chosen for the ESD generator is a capacity of 150 pF, which is representative of that of the human body. Furthermore, a discharge resistance of 330 Ω represents the source resistance of the human body when holding a metallic object in hand (such as a key or a tool). Figure K.26 shows the simplified circuit of an ESD generator. In the case of an air discharge test, the discharge switch, which is used for contact discharge, shall be closed.

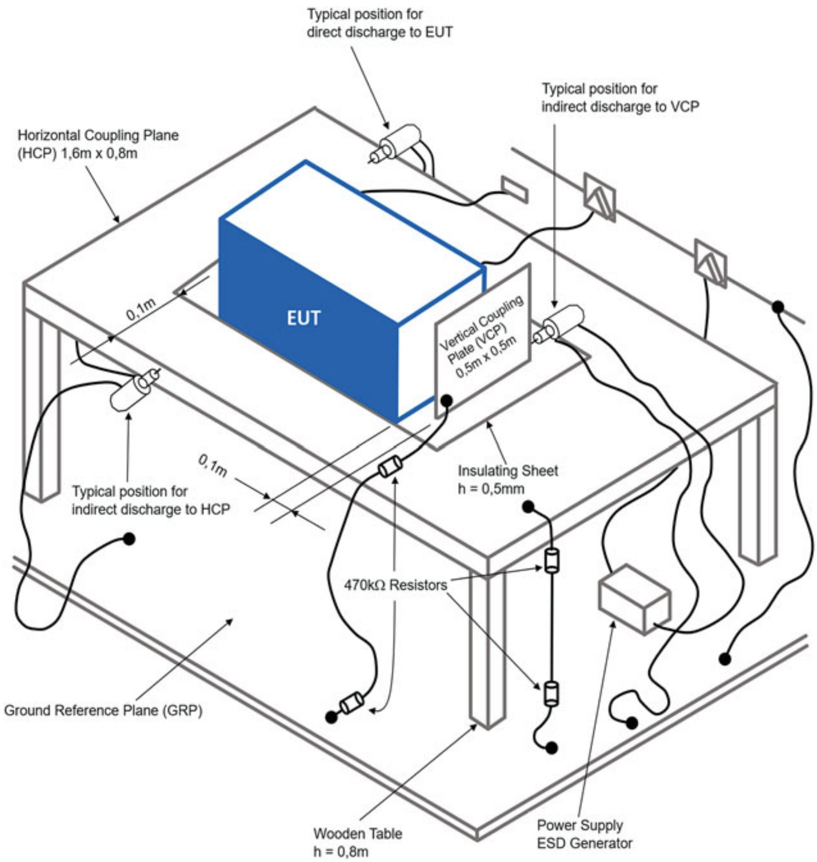


Fig. K.24 ESD tabletop devices test setup [8]

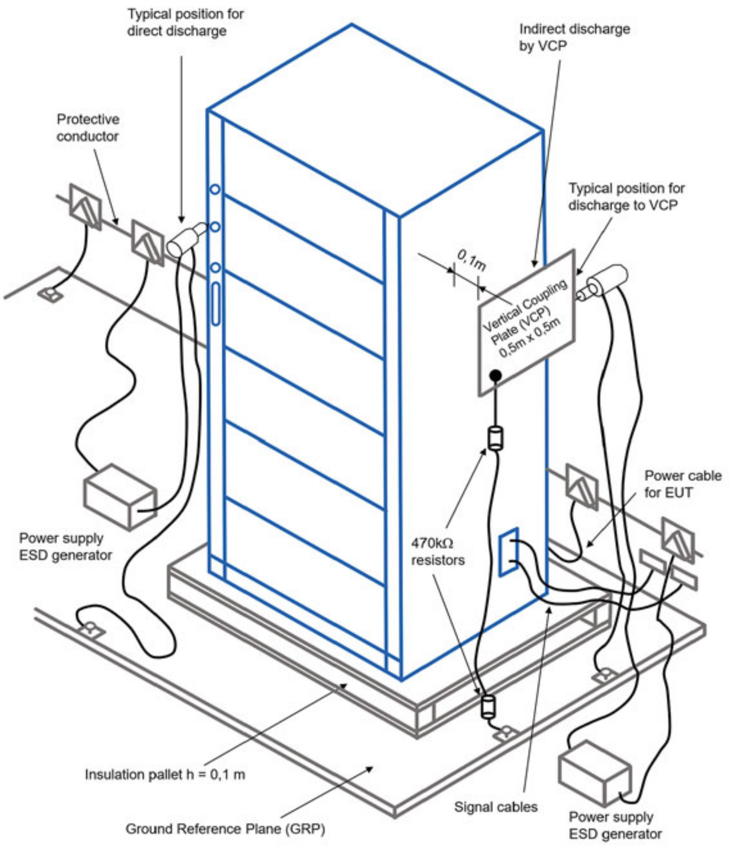


Fig. K.25 ESD tabletop devices test setup [8]

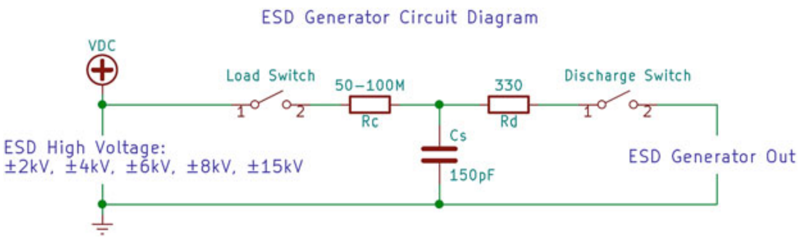


Fig. K.26 ESD test generator [8]

K.6.2 IEC 61000-4-2: Test Levels

The voltage levels are increased gradually until the maximum severity level selected is reached. Discharges to the EUT (Fig. K.27) and coupling planes are performed at a minimum of 1 sec intervals for each polarity (Tables K.5 and K.6).

Table K.7 helps determine which test level to apply for specific relative humidity levels and materials involved. Test levels are specified in the corresponding Product (Family) EMC Standard or Generic EMC Standard.

K.7 RF Radiated Immunity: IEC 61000-4-3

The IEC/EN 61000-4-3 EMC standard is about immunity testing (measurement techniques) against radiated, radio-frequency electromagnetic fields. This standard is part of the Basic EMC Publications.

The object of IEC 61000-4-3 is to establish a common reference for evaluating the immunity of electrical and electronic equipment when subjected to radiated, radio-frequency electromagnetic fields. The test method documented in this part of IEC 61000 describes a consistent method to assess the immunity of an equipment or

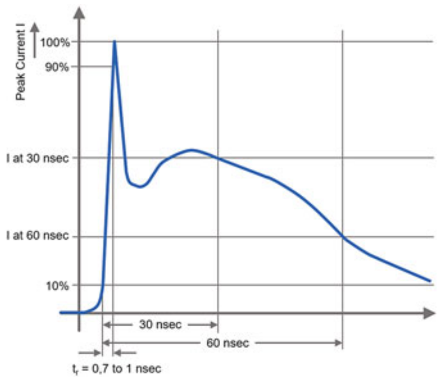


Fig. K.27 ESD discharge current waveform [8]

Table K.5 ESD discharge current waveform parameters [8]

Test Level	Discharge Current Waveform Parameter				
	Displayed Voltage	First Discharge Current Peak (± 10 %)	Risetime t_r with discharge switch	Current at 30 nsec (± 30 %)	Current at 60 nsec (± 30 %)
1	2 kV	7.5 A	0.7 to 1 nsec	4	2
2	4 kV	15 A	0.7 to 1 nsec	8	4
3	6 kV	22.5 A	0.7 to 1 nsec	12	6
4	8 kV	30 A	0.7 to 1 nsec	16	8

Waveform of the output current of the ESD test generator must match the required waveform.

Table K.6 ESD test voltage levels [8]

Test Level	Test Voltage	
	Contact Discharge	Air Discharge
1	2 kV	2 kV
2	4 kV	4 kV
3	6 kV	8 kV
4	8 kV	15 kV
X[a]	Special	Special

[a] X is an open test level
Testing shall also be satisfied at all lower levels given in the table

Table K.7 Instructions for selecting the ESD immunity test level [8]

Test Level	Instructions for selecting the ESD immunity test level			
	Relative humidity, as low as described below	Anti-static material	Synthetic material	Highest voltage
1	35%	X		2 kV
2	10%	X		4 kV
3	50%		X	8 kV
4	10%		X	15 kV

system against a defined phenomenon. This part deals with immunity tests related to the protection against RF electromagnetic fields from any source. Particular considerations are devoted to the protection against radio-frequency emissions from digital radiotelephones and other RF emitting devices.

K.7.1 IEC 61000-4-3: Test Setup

Tests have to be conducted in a shielded anechoic chamber (due to the high field strength) (Fig. K.28). The distance between the RF generation antenna and the equipment under test (EUT) is typically 3 m.Testing should be performed in a configuration as close as possible to the actual conditions in which the EUT will be used. A metallic grounding plane is not required, but the EUT should be placed on a table or support made of a non-metallic, non-conductive material (0.8 m table height for tabletop equipment, 0.05 m to 0.15 m support height for floor-standing equipment). Low dielectric constant materials such as polystyrene are recommended for non-conducting tables and supports. Materials like wood can become reflective when exposed to higher frequencies.

Here is a list of the most essential test equipment needed [9]:

- **Anechoic chamber.** The dimensions of the anechoic chamber must be sufficient to produce a homogeneous field of sufficient size with respect to the EUT. Additional absorbers can be used to reduce reflections in chambers that are not fully lined.
- **EMI Filters.** Not always necessary. Filters should not add any additional resonance effects during the test.

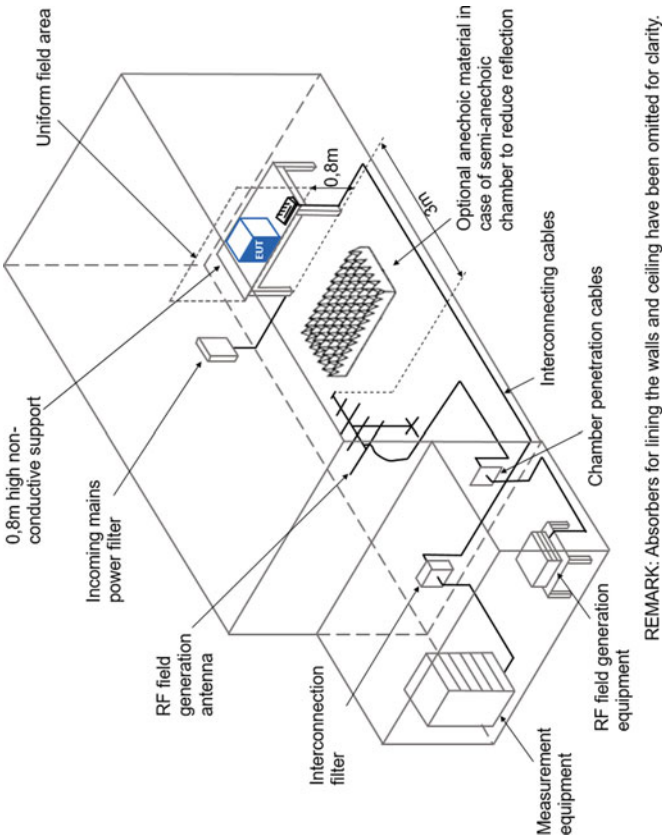


Fig. K.28 IEC 61000-4-3 RF immunity test setup [9]

- **RF signal generator.** The RF signal generator must be able to generate in the frequency band of interest an amplitude modulated signal (1 kHz sine wave with 80% modulation depth).
- **Power amplifier.** The power amplifier enables that the antenna can emit the required field strength. Distortion: Harmonics must be at least 6dB below the fundamental frequency. Before testing, a linearity check must be performed to ensure that the RF amplifier is not operating in compression.
- **Field generating antenna.** These can be either bi-conical, log-periodic, horn, or any other linearly polarized antenna system that fulfills the frequency requirements.
- **Isotropic field sensor.** The sensor must have adequate immunity to the field strength being measured. Usage of a fiber-optic link to an indicator outside the chamber is recommended.
- **Equipment to record power levels.** This equipment is for logging the necessary power levels to generate the required field strength.

K.7.2 IEC 61000-4-3: Test Levels

IEC 61000-4-3 does not suggest that a single test level is applicable over the entire frequency range. Rather more, it should be tested for the appropriate test level for each frequency range (Table K.8).

Testing for general-purpose RF immunity covers the 80 MHz to 1000 MHz frequency range and should be performed without any gaps.

For mobile communication and other higher frequency devices, tests should be performed in the 800 MHz to 960 MHz and 1.4 GHz to 6.0 GHz frequency ranges. Tests in these ranges do not need to be applied continuously over the entire range, and the ranges may be limited to specific frequencies for compliance with specific operating bands in the country the product will be sold in.

The specific product standards establish the most appropriate level for each frequency range. The test levels and the frequency bands are selected in accordance with the electromagnetic radiation environment to which the EUT can be exposed when finally installed. The electromagnetic environment for radiated immunity according to IEC 61000-4-3 is divided into the following classes [9]:

- **Class 1.** Low-level electromagnetic radiation environment. Levels typical of local radio/television stations located at more than 1 km and transmitters/receivers of low power.
- **Class 2.** Moderate electromagnetic radiation environment. Low power portable transceivers (typically less than 1 W rating) are in use, but with restrictions on use in close proximity to the equipment. A typical commercial environment.
- **Class 3.** Severe electromagnetic radiation environment. Portable transceivers (2 W rating or more, cell phones' peak power is 2 W [4]) are in use relatively close but not less than 1 m. High power broadcast transmitters are in close proximity and ISM equipment may be located close by. A typical industrial environment.
- **Class 4.** Portable transceivers are in use within less than 1 m of the equipment. Other sources of significant interference may be within 1 m of the equipment.
- **Class X:** X is an open level which might be negotiated and specified in the product standard or equipment specification.

Table K.8 IEC 61000-4-3 RF immunity test levels [9]

Test Level	Field Strength
1	1 V/m
2	3 V/m
3	10 V/m
4	30 V/m
X[a]	Special

[a] X is an open test level and the associated field strength can be of any value. This test level can be specified in the EMC Product Standard.

K.8 EFT: IEC 61000-4-4

The IEC/EN 61000-4-4 EMC standard is about immunity testing (measurement techniques) against repetitive electrical fast transients (EFT), also called bursts. This standard is part of the Basic EMC Publications.

IEC 61000-4-4 defines immunity requirements, test setups, test procedures, test equipment (and their calibration and verification), and ranges of test levels related to repetitive electrical fast transients (EFTs, bursts). In reality, bursts originate from switching transients: interruption of inductive loads, relay contact bounce, etc.

The object of IEC 61000-4-4 is to establish a common and reproducible reference in order to evaluate the immunity when subjected to bursts on the following ports [10]:

- Supply ports
- IO (signal, control) ports (if cable length > 3 m)
- Earth ports

K.8.1 IEC 61000-4-4: Test Setup

Floor standing EUTs and equipment designed to be mounted in other configurations, unless otherwise mentioned, shall be placed on a ground reference plane and shall be insulated from it with an insulating support with a thickness of $0.1\,\mathrm{m} \pm 0.05\,\mathrm{m}$ including non-conductive castors. Tabletop equipment and equipment normally mounted on ceilings or walls as well as built-in equipment shall be tested with the EUT located $0.1\,\mathrm{m} \pm 0.01\,\mathrm{m}$ above the ground reference plane. Figure K.29 shows locations for supply line coupling (A) and location for signal lines coupling (B).

The simplified circuit diagram of the generator is given in Fig. K.30. The effective output impedance of the generator shall be $50\,\Omega$. Here is the explanation of the components:

- R_C. Charging resistor
- C_C. Energy storage capacitor
- R_S. Impulse duration shaping resistor
- R_m. Impedance matching resistor
- C_d. DC blocking capacitor $10\,\mathrm{nF} \pm 2\,\mathrm{nF}$

The burst generator output signal consists of periodic burst pulse packages (75 pulses per package). The burst period is 300msec and the repetition period is either $10\,\mu\mathrm{s}$ (100 kHz) or $200\,\mu\mathrm{s}$ (5 kHz) (Fig. K.31).

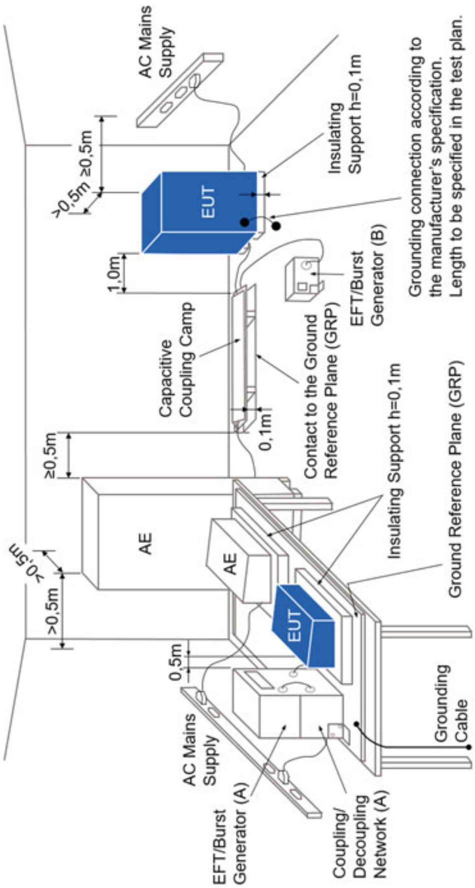


Fig. K.29 IEC 61000-4-4 EFT/bursts test setup [10]

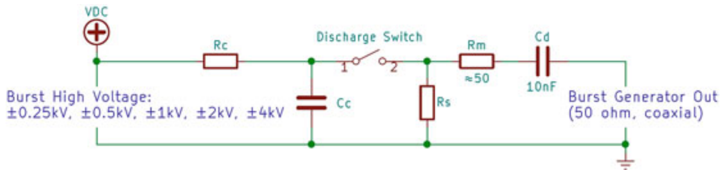


Fig. K.30 EFT voltage bursts test generator [10]

K.8.2 IEC 61000-4-4: Test Levels

The use of 5 kHz repetition frequency is traditional (Table K.9). However, 100 kHz is closer to reality. Product committees should determine which frequencies are relevant for specific products or product types. For example, with some products, there may be no clear distinction between power ports and signal ports, in which case it is up to product committees to make this determination for test purposes.

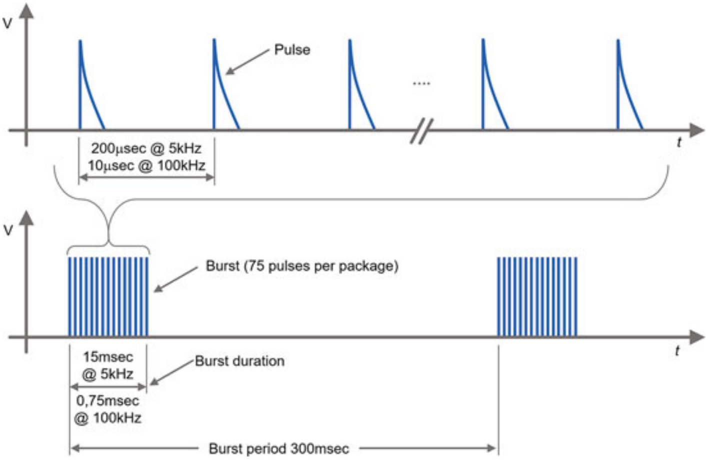


Fig. K.31 EFT burst pulse packages [10]

Table K.9 IEC 61000-4-4 EFT (bursts) voltage test levels [10]

Open circuit output test voltage and repetition frequency of the impulses				
	Power ports, earth ports (PE)		Signal and control ports	
Test Level	Voltage peak [kV]	Repetition frequency [kHz]	Voltage peak [kV]	Repetition frequency [kHz]
1	0.5	5 or 100	0.25	5 or 100
2	1	5 or 100	0.5	5 or 100
3	2	5 or 100	1	5 or 100
4	4	5 or 100	2	5 or 100
X[a]	Special	Special	Special	Special

[a] "X" can be any level, above, below or in between the others. The level shall be specified in the dedicated equipment specification

Here the respective electromagnetic environmental classes for the test levels above [10]:

- **Level 1.** Well-protected environment. A computer room may represent this environment.
- **Level 2**. Protected environment. The control room or terminal room of industrial and electrical plants may represent this environment.
- **Level 3.** Typical industrial environment. The area of industrial process equipment may represent this environment.
- **Level 4.** Severe industrial environment. The outdoor area of industrial process equipment where no specific installation practice has been adopted, power plants, the relay rooms of open-air high voltage substations, and gas insulated substations of up to 500kV operating voltage (with typical installation practice) may represent this environment.
- **Level X.** Special situations to be analyzed. The minor or major electromagnetic separation of disturbance sources from equipment circuits, cables, lines, etc., and the quality of the installations may require the use of a higher or lower environmental level than those described above. It should be noted that equipment lines of a higher environmental level can penetrate a lower severity environment.

K.9 Surge: IEC 61000-4-5

The IEC/EN 61000-4-5 EMC standard is about immunity testing (measurement techniques) against surges (high energy, high voltage pulses). This standard is part of the Basic EMC Publications.

IEC 61000-4-5 defines immunity requirements, test setups, test procedures, test equipment (and their calibration and verification), and ranges of test levels related to unidirectional surges caused by over-voltages from switching and lightning transients. Direct injections of lightning currents, i.e., direct lightning strikes, are not considered in this standard.

The object of IEC 61000-4-5 is to establish a common reference for evaluating the immunity of electrical and electronic equipment when subjected to surges.

K.9.1 IEC 61000-4-5: Test Setup

The test setup comprises the following equipment [13]:

- Equipment under test (EUT)
- Auxiliary/supporting equipment (AE)
- Cables of defined types and length
- Coupling network (usually: capacitive, in cases of high-bandwidth signal lines: arrestors)
- Test generator (combination wave generator, 700 μs generator)
- Decoupling network and protection devices
- Additional test generator source resistors (10 Ω and 40 Ω)

Two types of surge pulse generators are specified in IEC 61000-4-5.

- **10/700 μs.** The surge pulse generator for generating the pulse shape 10/700 μs is used to test ports intended for connection to outdoor symmetrical communication lines.
- **1.2/50 μs.** The surge pulse generator for generating the pulse shape 1.2/50 μs is used in all other cases, particularly for testing ports to power supply lines.

The simplified circuit diagrams of the 1.2/50 μs and the 10/700 μs hybrid surge generators are given in Figs. K.32 and K.33. The effective output impedance shall be 2 Ω ± 10% for the 1.2/50 μs generator and 40 Ω ± 10% for the 10/700 μs generator. Here are the explanation of the components:

- R_C. Charging resistor.
- C_C. Energy storage capacitor.
- R_S. Impulse duration shaping resistors.
- R_m. Impedance matching resistors.
- L_r. Rising time forming inductor (1.2/50 μs).
- C_s. Impulse duration shaping capacitor (0.2 μF, 10/700 μs only).

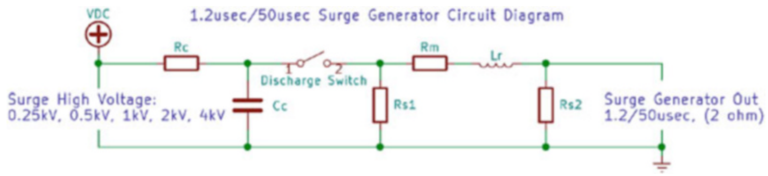


Fig. K.32 1.2/50 µs-surge voltage test generator [13]

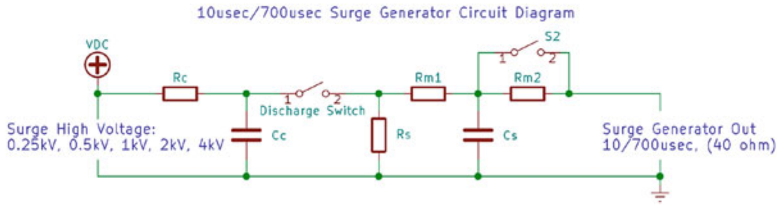


Fig. K.33 10/700 µs-surge voltage test generator [13]

- S_2. Switch (closed if external matching resistors are attached, 10/700 µs only).

The selection of the output (or source) impedance of the surge generator depends on:

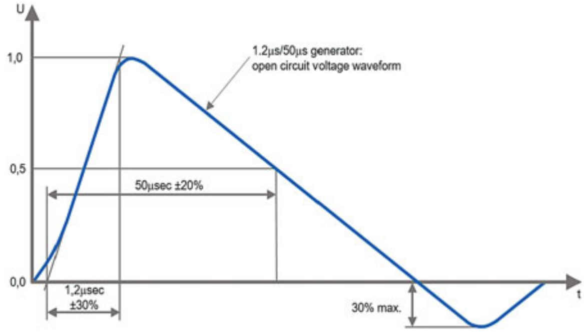
- Type of cables/conductors (AC power lines, DC power lines, signal lines, etc.).
- The length of the cables and lines.
- Conditions inside and outside buildings.
- Coupling of the test voltage (between lines or between line and earth).

Some countries (e.g., the USA) refer to other non-IEC standards that specify lower source impedances (more stringent testing). The EMC standard IEC 61000-4-5 defines the following source impedances [13]:

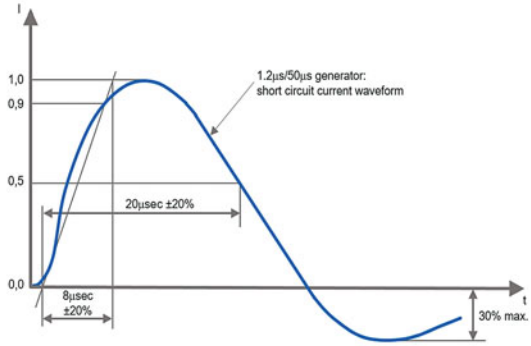
- **Public mains line to line.** $2\,\Omega$ ($R_{ext} = 0\,\Omega + R_{out} = 2\,\Omega$) represents the source impedance of the low-voltage power supply network (public mains).
- **Public mains line to earth.** $12\,\Omega$ ($R_{ext} = 10\,\Omega + R_{out} = 2\,\Omega$) represents the source impedance of the low-voltage power supply network and ground (common mode).
- **Signal lines to earth.** $42\,\Omega$ ($R_{ext} = 40\,\Omega + R_{out} = 2\,\Omega$) represents the source impedance between all other lines and ground

Figures K.34 and K.35 show the 1.2/50 µs generator and the 10/700 µs generator open circuit voltage and short circuit current waveforms.

Fig. K.34 1.2/50 μs-surge generator waveforms. (**a**) Open circuit voltage waveform [13]. (**b**) Short circuit current waveform [13]



(a)



(b)

K.9.2 IEC 61000-4-5: Test Levels

Table K.10 shows the defined test levels of IEC 61000-4-5. Devices and installations connected to the public mains must have the minimum immunity level of:

- Coupling between lines: 0.5 kV
- Coupling between lines to earth: 1 kV

The maximum peak current for each test level can be found in Table K.11: The selection of the test level should be based on the installation conditions. Unless otherwise specified in the product and product family standard, the table below should be used for this purpose. Here are the defined installation classes of IEC 61000-4-5 [13]:

- **Class 0.** Well-protected electrical environment, often in a special room.
- **Class 1.** Partially protected electrical environment.
- **Class 2.** Electrical environment where the cables are well separated, even at short runs.
- **Class 3.** Electrical environment where power and signal cables run in parallel.

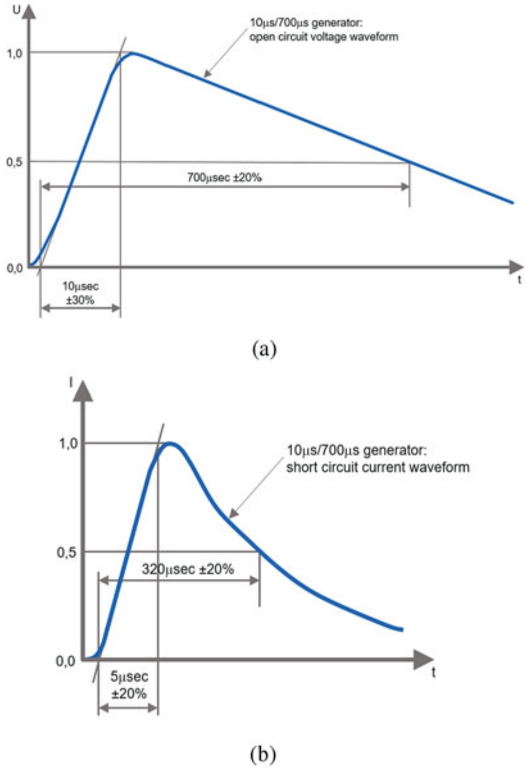


Fig. K.35 10/700 µs-surge generator waveforms. (**a**) Open circuit voltage waveform [13]. (**b**) Short circuit current waveform [13]

Table K.10 IEC 61000-4-5 surge voltage test levels [13]

Open circuit output test voltage	
Test Level	**Voltage peak ±10%** [kV]
1	0.5
2	1
3	2
4	4
X[a]	Special

[a] "X" can be any level, above, below or in between the others. The level shall be specified in the dedicated equipment specification

- **Class 4.** Electrical environment where the interconnections include outdoor cables along with the power cable, and cables are used for both electronics and electric circuits.
- **Class 5.** Electrical environment for electronic equipment connected to telecommunication cables and overhead power lines in a non-densely populated area.
- **Class X.** Special conditions specified in the product specification.

Table K.11 IEC 61000-4-5 surge test maximum currents [13]

Maximum peak current (±10%) values depending on voltage level and R_{ext}					
$R_{out} + R_{ext}$	Test Level 1 500 V	Test Level 2 1 kV	Test Level 3 2 kV	Test Level 4 4kV	Examles
2 Ω	250 A	500 A	1 kA	2 kA	AC (DC) power: line-to-line
12 Ω	42 A	84 A	167 A	334 A	AC (DC) power: line-to-earth
42 Ω	12 A	24 A	48 A	96 A	Signals: line-to-line, line-to-earth

K.10 RF Conducted Immunity: IEC 61000-4-6

The IEC/EN 61000-4-6 EMC standard is about immunity testing (measurement techniques) against conducted disturbances induced by radio-frequency fields (RF, 9 kHz to 80 MHz). This standard is part of the Basic EMC Publications. Equipment not having at least one conducting wire and/or cable (such as mains supply, signal line, or earth connection) which can couple the equipment to the disturbing RF fields is excluded from the scope of this publication. No tests are required in the 9 kHz–150 kHz range.

The objective of IEC 61000-4-6 is to establish a common reference for evaluating the functional immunity of electrical and electronic equipment when subjected to conducted disturbances induced by RF fields. Typical RF transmitters (sources of emissions) are:

- Transmitting radio systems (e.g., radio, television, mobile phones, wireless phones) cause fields which induce disturbances in lines.
- Low-frequency interference currents of power electronics (e.g., power converters or motor drivers).

It is assumed that electromagnetic field disturbances may act on the whole length of cables connected to the installed equipment. The dimensions of the disturbed equipment are assumed to be small compared to the wavelengths involved. The in-going and out-going cables and wires (e.g., mains, communication lines, interface cables) behave as passive receiving antenna networks (they can be several wavelengths long). The susceptible equipment is exposed to currents flowing through the equipment between those cable networks. Cable systems connected to equipment are assumed to be in resonant mode ($\lambda/4$, $\lambda/2$ dipoles) and are represented by coupling and decoupling devices having a common-mode impedance of 150 Ω with respect to a ground reference plane.

K.10.1 IEC 61000-4-6: Test Setup

The effect of conducted disturbing signals, induced by electromagnetic radiation, is tested by injecting the signal via particular Coupling/Decoupling Networks (CDNs) to the cabling. All cables exiting the EUT shall be supported at a height of at least 30 mm above the reference ground plane (Figs. K.36 and K.37).

The IEC 61000-4-6 test method subjects the EUT to a source of disturbance comprising electric and magnetic fields, simulating those coming from intentional RF transmitters. The disturbing fields (E [V/m] and H [A/m]) are approximated by the electric and magnetic near-fields resulting from the voltages and currents caused by the test setup. The use of coupling and decoupling devices to apply the disturbing signals to one cable at a time, while keeping all other cables non-excited, can only approximate the real situation where disturbing sources act on all cables simultaneously, with a range of different amplitudes and phases.

The appropriate coupling of the disturbing signal to the cables connected to the EUT can be achieved by coupling and decoupling devices. Here are the different types of coupling devices:

- **Direct injection devices.** Signal coming from the test generator is injected onto shielded and coaxial cables via a 100 Ω resistor. In between the Auxiliary Equipment (AE) and the injection point, a decoupling network (like mentioned below) should be inserted as close as possible to the injection point.
- **Coupling/decoupling networks (CDNs).** CDNs are the preferred coupling devices because of their test reproducibility and built-in protection of the Auxiliary Equipment (AE). These networks combine the coupling and decoupling circuits in a housing, e.g., CDN-M1, CDN-M2, DCN-M3, CDN-T2, CDN-T4, and CDN-AF-2. There are different series of CDNs, depending on the cable signal type:
  - M: Power line/mains connection.
  - C: Shielded coaxial cable.
  - T: Unshielded balanced signal pairs.
  - AF: Unshielded and unbalanced signal cables.
  - S: Shielded signal cables.
- **Clamp injection devices.** Coupling and decoupling functions are separated, coupling is provided by a clamp on device, while common mode impedance and decoupling functions are established at the Auxiliary Equipment (AE).
  - **Current clamps.** Establish an inductive coupling to the cable connected to the EUT.
  - **EM clamps.** Establish a capacitive and an inductive coupling to the cable connected to the EUT.
- **Decoupling networks.** Decoupling networks typically includes multiple inductors to create a high common-mode impedance over the frequency range of interest. This is determined by the ferrite material used. The inductance has to be at least 280 μH at 150 kHz. The effective resistance must remain high: $\geq$ 260 Ω to 26 MHz and $\geq$ 150 Ω above 26 MHz.

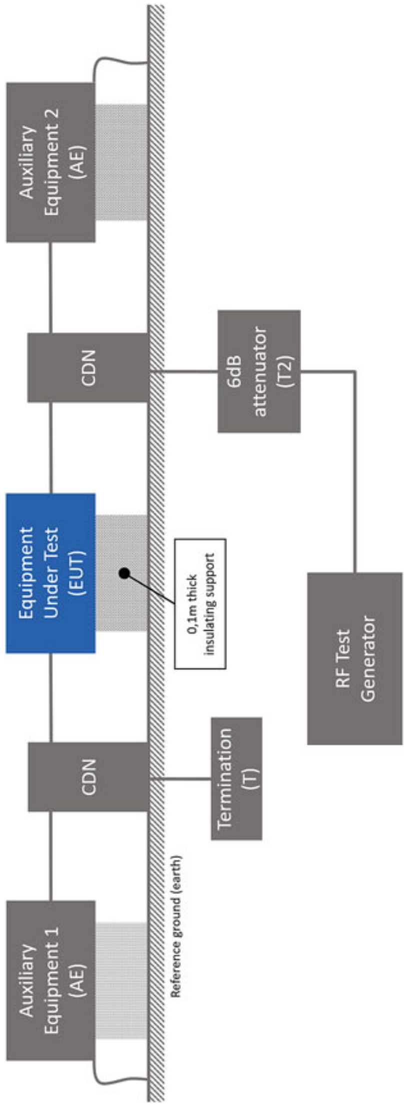


Fig. K.36 IEC 61000-4-6 conducted RF immunity test setup [11]

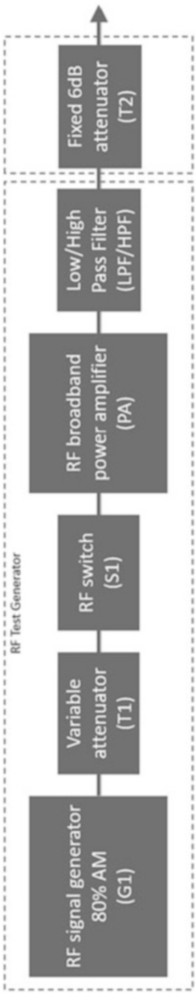


Fig. K.37 IEC 61000-4-6 RF signal test generator [11]

Table K.12 IEC 61000-4-6 conducted RF immunity test levels [11]

Frequency Range 150 kHz to 80 MHz		
Test Level	**Voltage V_0 (EMF)** [V/m]	
	[dBuV]	**[V]**[b]
1	120	1
2	130	3
3	140	10
X[a]	Special	

[a] X is an open test level

[b] Root mean square (RMS) value of the unmodulated EMF signal

K.10.2 IEC 61000-4-6: Test Levels

The EUT is subjected to an electromotive force (EMF) of 1 V, 3 V, or 10 V from 150 kHz to 80 MHz (Table K.12). This frequency range is 80% amplitude modulated (AM) with a 1 kHz sine wave. The RF signal generator provides the modulated frequency at a step rate of 1% of fundamental to the RF signal. The selection of the test level depends on the electromagnetic environment where the EUT will be used in. The immunity test levels of IEC 61000-4-6 are divided into the classes below [11]:

- **Class 1.** Low-level electromagnetic radiation environment. Typical level where radio and/or television stations are located at a distance of $>$ 1 km and typical level for low power transceivers.
- **Class 2.** Moderate electromagnetic radiation environment. Low power portable transceivers (typically $<$ 1 W rating) are in use, but with restrictions on use in close proximity to the equipment. A typical commercial environment.
- **Class 3.** Severe electromagnetic radiation environment. Portable transceivers (2 W rating or more, cell phones' peak power is 2 W [4]) are in use relatively close to the equipment, but at a distance not less than 1 m. High-powered broadcast transmitters are in close proximity to the equipment and ISM equipment may be located close by. A typical industrial environment.
- **Class X.** X is an open immunity test level that can be agreed and recorded in the respective product standard or device specification.

K.11 Magnetic Field Immunity: IEC 61000-4-8

The IEC/EN 61000-4-8 relates to the immunity requirements of equipment, only under operational conditions, to radiated magnetic disturbances at power frequencies 50 Hz and 60 Hz. This standard is part of the Basic EMC Publications (according to IEC Guide 107). IEC 61000-4-8 defines:

- Immunity test levels
- Test equipment
- Test setup
- Test procedure

IEC 61000-4-8 does not consider disturbances due to capacitive or inductive coupling in cables or other parts of the installation. However, other IEC standards dealing with conducted disturbances cover these aspects. The object of IEC 61000-4-8 is to establish a common and reproducible basis for evaluating the performance of electrical and electronic equipment for household, commercial, and industrial applications when subjected to magnetic fields at power frequency (continuous and short-duration field).

In reality, the source of the magnetic fields is typically the power line current flowing in conductors, or occasionally, transformers in nearby equipment. IEC 61000-4-8 distinguishes between the following two cases of interference due to low-frequency magnetic fields (50 Hz, 60 Hz):

- The source of interference produces (under normal operating conditions) a continuous (steady-state) magnetic field of comparably small amplitude.
- The source of interference produces (in case of an error or failure) a short-term (pulsed) magnetic field of comparably high amplitude. This magnetic field is of short duration, only until the protective elements respond (a few milliseconds for fuses, a few seconds for protective relays).

Other types of magnetic fields are subject to further standards:

- Fields with other energy-related frequencies (16.7 Hz, 20 Hz, 30 Hz, 400 Hz)
- Fields with harmonic currents (100 Hz to 2000 Hz)
- Fields with higher frequencies (up to 150 kHz, e.g., for signal transmission on low-voltage electrical networks)
- DC static fields

K.11.1 IEC 61000-4-8: Applicability

IEC 61000-4-8 applies to equipment used at different locations [12]:

- Residential and commercial locations
- Near industrial installations and power plants
- Near medium voltage and high-voltage sub-stations

The applicability of IEC 61000-4-8 to equipment installed in these different locations is determined by the presence of the following phenomenon:

- The continuous magnetic field test is applicable to all equipment intended for public or industrial low-voltage networks or electrical installations.

- The short-term (pulsed) magnetic field test is typically applicable to equipment installed in places with a particularly harsh electromagnetic environment.

K.11.2 IEC 61000-4-8: Test Setup

The test setup is simple and consists of the following parts:

- Equipment Under Test (EUT).
- Ground reference plane, non-magnetic (copper or aluminum).
- Test generator.
- Induction coil, non-magnetic (copper or aluminum).

The method for testing the immunity to magnetic fields is to produce a controlled magnetic field of known field strength by driving a large coil with a test generator, and placing the equipment in the center of the coil, thereby subjecting the equipment to the magnetic field. Figure K.38 shows the principle circuit diagram of a test generator: One or several induction coils generate the magnetic field. Over the complete volume of the EUT, the magnetic field strength must not fluctuate more than ±3 dB. The EUT dimensions must be equal or smaller than the testing volume. Below is a list of typical loop coil test antennas. However, other designs are possible.

- **Single square loop coil.** For example, for tabletop devices.
  Coil dimensions = 1 m × 1 m.
  Testing volume = 0.6 m × 0.6 m × 0.5 m high.
  Minimum spacing between EUT and coil = 0.2 m.
- **Double square loop coils.** For example, for floor standing devices.
  Coil dimensions = 1 m × 1 m. Coils are 0.6 m spaced.
  Testing volume = 0.6 m × 0.6 m × 1 m high (0.8 m hight for 0.8 m spacing).
  Minimum spacing between EUT and coil = 0.2 m.
- **Single rectangular loop coil.** For example, for floor standing devices.
  Coil dimension = 1 m × 2.6 m.
  Testing volume = 0.6 m × 0.6 m × 2 m high.
  Minimum spacing between EUT and coil = 0.2 m from and 0.3 m from shorter side.

Testing takes place in three orthogonal orientations (by rotating the antenna by 90°). The EUT has to be placed on a 10 cm insulating support (e.g., dry wood) over the ground reference plane. Some example test setups are shown in Fig. K.39.

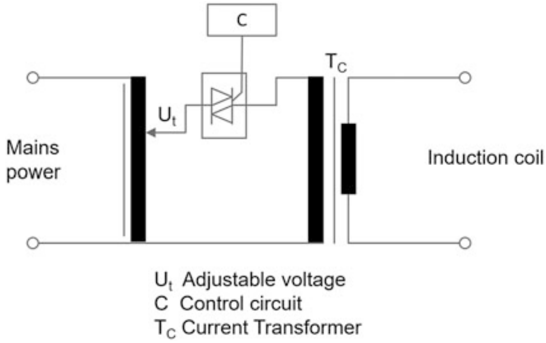


Fig. K.38 IEC 61000-4-8 test generator [12]

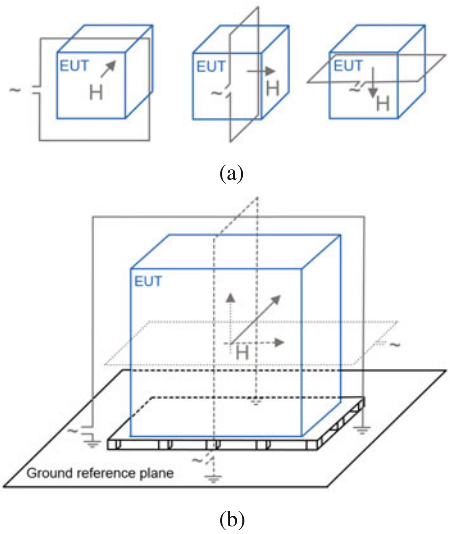


Fig. K.39 IEC 61000-4-8 magnetic field immunity test setup. (**a**) Small EUTs [12]. (**b**) Large EUTs [12]

K.11.3 IEC 61000-4-8: Test Levels

The required immunity test level for a specific product is specified in the corresponding EMC product standard, EMC family standard, or the generic EMC standard (Table K.13). The immunity test levels are specified in [A/m]. In free space, a magnetic field strength of 1 A/m corresponds to a magnetic flux density of 1.26 μT. The selection of a test level depends on the expected operating environment of the equipment under test (EUT). Therefore, an appropriate test level for an EUT should be chosen based upon the magnetic field strengths the EUT is likely to encounter in its typical operating environment. The environments in which equipment can operate are divided into five classes:

Table K.13 IEC 61000-4-8 magnetic field immunity test levels [12]

Test Level	Continuous field strength	Pulsed (1 to 3sec) field strength
1	1 A/m	Not applicable
2	3 A/m	Not applicable
3	10 A/m	Not applicable
4	30 A/m	300 A/m
5	100 A/m	1000 A/m
X	Special	Special

X is an open test level. This immunity test level can be specified in the EMC product standard.

- **Class 1.** Environmental level where sensitive device using electron beam can be used. Examples include environments containing CRT monitors or an electron microscope.
- **Class 2.** Well-protected environment. Examples include household, office, and hospitals.
- **Class 3.** Protected environment. Examples include commercial areas, small industrial plants, or a computer room of a high-voltage sub-station.
- **Class 4.** Industrial environment. Examples include heavy industrial plants, power plants, or the control room of a high-voltage sub-station.
- **Class 5.** Severe industrial environment. Examples include the switchyard of heavy industrial plants or medium-voltage and high-voltage power stations.

K.12 AC Dips: IEC 61000-4-11

The IEC/EN 61000-4-11 EMC standard is about immunity testing (measurement techniques) of voltage dips, short interruptions, and voltage variations. This standard is part of the Basic EMC Publications (according to IEC Guide 107). IEC 61000-4-11 defines the immunity test methods and test levels for electrical and electronic equipment connected to low-voltage power supply networks for:

- **Voltage dips.** A sudden reduction of the AC supply voltage (at any phase angle) below a specified dip threshold, followed by its recovery after a brief interval.
- **Short interruptions.** A sudden reduction of the AC supply voltage on all phases (at an any phase angle) below a specified interruption threshold, followed by its restoration after a brief interval. Short interruptions can be considered as voltage dips to 0V.
- **Voltage variations.** Gradual changes of AC voltage to a higher or lower value than the nominal voltage value. The duration can be short or long.

Voltage dips or short interruptions are typically caused by faults in the public mains network. Voltage variations are typically the result of varying loads connected to the public mains.

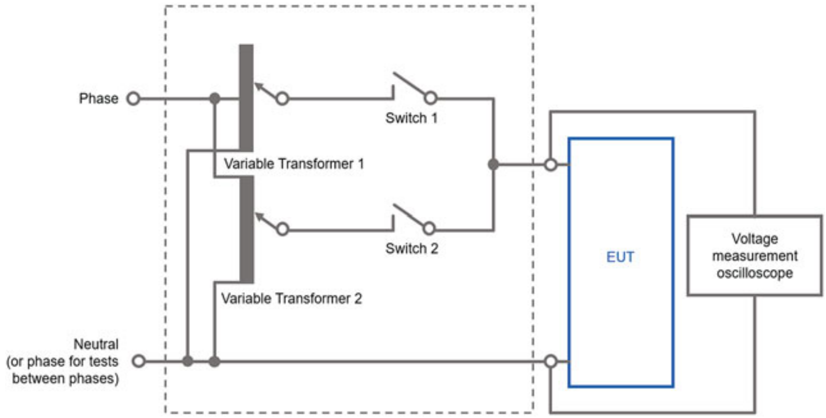


Fig. K.40 IEC 61000-4-11 test setup example [7]

K.12.1 IEC 61000-4-11: Applicability

This standard applies to electrical and electronic equipment having a rated input current not exceeding 16A per phase for connection to 50 Hz or 60 Hz AC networks. It does not apply to electrical and electronic equipment for connection to DC or 400 Hz AC networks.

K.12.2 IEC 61000-4-11: Test Setup

The equipment under test (EUT) must be connected with the shortest specified power cable of the EUT. Figure K.40 shows one of several possible test setups.

K.12.3 IEC 61000-4-11: Test Levels

The test level voltages in IEC 61000-4-11 are stated relative to the rated equipment voltage (UT). The change between UT and the changed voltage is abrupt. The step can start and stop at any phase angle on the mains voltage. The following test voltage levels (in % UT) are used: 0%, 40%, 70%, and 80%, corresponding to dips with residual voltages of 0%, 40%, 70%, and 80%.

The required immunity test level for a specific product is specified in the corresponding EMC product standard, EMC family standard or in the generic EMC standard. The selection of a test level depends on the expected operating environment of the equipment under test (EUT). The environments in which equipment can operate are divided into three classes [7]:

- **Class 1.** This class applies to protected supplies and has compatibility levels lower than public network levels. It relates to the use of equipment very sensitive to disturbances in the power supply, for instance, the instrumentation of technological laboratories, some automation and protection equipment, some computers, etc. Class 1 environments normally contain equipment which requires protection by such apparatus as uninterruptible power supplies (UPS), filters, or surge suppressors.
- **Class 2.** This class applies to points of common coupling (PCC's for consumer systems) and in-plant points of common coupling (IPC's) in the industrial environment in general. The compatibility levels in this class are identical to those of public networks; therefore components designed for application in public networks may be used in this class of industrial environment.
- **Class 3.** This class applies only to in-plant points of common coupling IPC's in industrial environments. It has higher compatibility levels than those of class 2 for some disturbance phenomena. For instance, this class should be considered when any of the following conditions are met:
  - a major part of the load is fed through converters
  - welding machines are present
  - large motors are frequently started
  - loads vary rapidly.

The test levels for voltage dips and interruptions are given in the tables below. The preferred test levels and durations given in Tables K.14 and K.15 consider the information given in IEC 61000-2-8. The test levels for voltage variations are only optional. Therefore, they are not mentioned here. Some example graphs of voltage dips and short interruptions are shown in Figures K.41, K.42, and K.43.

Table K.14 IEC 61000-4-11 AC voltage dips test levels [7]

Preferred test levels and durations for voltage dips.					
Class	**Immunity test level and duration for voltage dips (t_s) for 50Hz/60Hz**				
1	Case-by-case according to the EUT requirements.				
2	0% during 0.5 cycle	0% during 1 cycle	70% for 25/30 cycles		
3	0% during 0.5 cycle	0% during 1 cycle	40% during 10/12 cycles	70% during 25/30 cycles	80% during 250/300 cycles
X	X	X	X	X	X

- Classes: Classes as per IEC 61000-2-4; see Annex B
- Class X: To be defined by product committee. For equipment connected directly or indirectly to the public network, the levels must not be less severe than Class 2
- Example: "25/30 cycles" means "25 cycles for 50 Hz test" and "30 cycles for 60 Hz test"

Table K.15 IEC 61000-4-11 AC voltage interruptions test levels [7]

Preferred test levels and durations for short interruptions.	
Class	**Immunity test level and duration for short interruptions (t_s) for 50Hz/60Hz**
1	Case-by-case according to the EUT requirements.
2	0% during 250/300 cycles
3	0% during 250/300 cycles
X	X

- Classes: Classes as per IEC 61000-2-4; see Annex B
- Class X: To be defined by product committee. For equipment connected directly or indirectly to the public network, the levels must not be less severe than Class 2
- Example: "250/300 cycles" means "250 cycles for 50 Hz test" and "300 cycles for 60 Hz test"

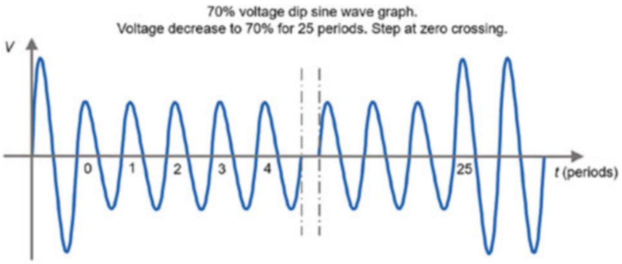


Fig. K.41 IEC 61000-4-11 70% voltage dip for 25 periods

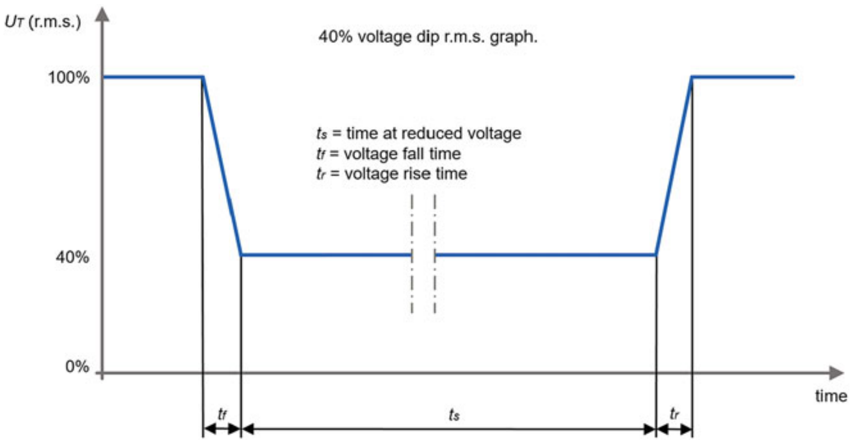


Fig. K.42 IEC 61000-4-11 40% voltage dip RMS

K.13 Generic Immunity Levels for Residential Environments: IEC 61000-6-1

The IEC 61000-6-1 is a Generic Standard for immunity (conducted, radiated) in residential, commercial, and light-industrial environments (indoor and outdoor). The immunity requirements cover the frequency range 0 Hz to 400 GHz.

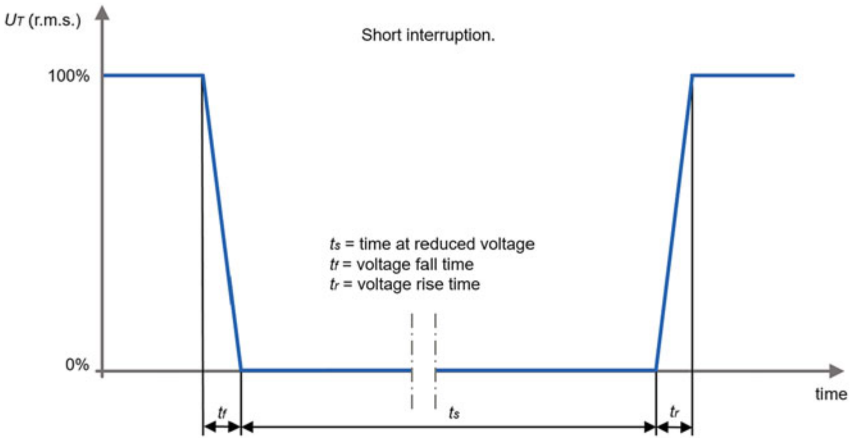


Fig. K.43 IEC 61000-4-11 short voltage interruption

IEC 61000-6-1 applies to:

- Equipment for which no relevant dedicated product standard or product family emission standard exists.
- Equipment intended to operate in residential locations (indoors, outdoors) and commercial, public, and light-industrial locations (indoors, outdoors).

K.14 Generic Immunity Levels for Industrial Environment: IEC 61000-6-2

The IEC 61000-6-2 (Generic standards—immunity standard for industrial environments) is a generic standard for immunity (conducted, radiated) in industrial environments (indoor and outdoor). The immunity requirements cover the frequency range 0 Hz to 400 GHz.

IEC 61000-6-2 applies to:

- Equipment for which no relevant dedicated product standard or product family emission standard exists.
- Equipment intended to operate in industrial locations (indoors, outdoors) and connected to the public mains power network or a DC distribution network or a battery.

K.15 Generic Emission Limits for Residential Environments: IEC 61000-6-3

The IEC 61000-6-3 is a Generic Standard for emissions (conducted, radiated) in residential environments (indoor and outdoor). The emission requirements cover the frequency range 0Hz to 400GHz.

IEC 61000-6-3 applies to:

- Equipment for which no relevant dedicated product standard or product family Emission standard exists.
- IEC 61000-6-3 for emission requirements applies to electrical and electronic equipment intended for use at residential locations. IEC 61000-6-3 also applies to electrical and electronic equipment intended for use at other locations that do not fall within the scope of IEC 61000-6-8 or IEC 61000-6-4.
- The intention is that all equipment used in the residential, commercial, and light-industrial environments are covered by IEC 61000-6-3 or IEC 61000-6-8. If there is any doubt, the requirements in IEC 61000-6-3 apply.
- The generic EMC standard for emissions of equipment that runs in light-industrial locations is IEC 61000-6-8.

K.16 Generic Emission Limits for Industrial Environment: IEC 61000-6-4

The IEC 61000-6-4 is a Generic Standard for emissions (conducted, radiated) in industrial environments (indoor and outdoor). The emission requirements cover the frequency range 9 kHz to 400 GHz.

IEC 61000-6-4 applies to:

- IEC 61000-6-4 is applicable to equipment that does not fall within the scope of IEC 61000-6-3.
- Equipment for which no relevant dedicated product standard or product family emission standard exists.
- Equipment intended for use within the environment existing at industrial locations.

K.17 Generic Emission Limits for Light-Industrial Environment: IEC 61000-6-8

The IEC 61000-6-8 is a Generic Standard for emissions (conducted, radiated) in light-industrial environments (indoor and outdoor). The emission requirements cover the frequency range 0 Hz to 400 GHz.

IEC 61000-6-8 applies to:

- IEC 61000-6-8 is applicable to light-industrial equipment, which satisfies the following restrictions of use:
  - is defined as professional equipment
  - is professionally installed and maintained
  - is not intended to be used in residential locations.

  Otherwise, IEC 61000-6-3 applies to electrical and electronic equipment intended for use at commercial and light-industrial locations that do not satisfy these restrictions.
- Equipment for which no relevant dedicated product standard or product family emission standard exists.

Examples of commercial or light-industrial locations are:

- Retail outlets
- Business premises
- Areas of public entertainment
- Places of worship
- Outdoor locations
- General public locations
- Hospitals, educational institutions
- Public traffic area, railway stations, and public areas of an airport
- Specific common area of buildings, such as basements, control rooms, electrical service areas
- Workshops, laboratories, service centers

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Appendix L
American Wire Gauge (AWG)

One who makes no mistakes makes nothing at al.

—Giacomo Casanova

This chapter presents conversions of different wire diameter units like [mm] or [mm^2] to AWG (Table L.1).

Table L.1 American Wire Gauge (AWG)

AWG	Diameter [in]	Diameter [mm]	Area mm^2	$R'_{DC,Cu}$ [mΩ/m]	AWG	Diameter [in]	Diameter [mm]	Area mm^2	$R'_{DC,Cu}$ [mΩ/m]
1	0.289	7.35	42.4	0.42	26	0.0159	0.405	0.129	138
2	0.258	6.54	33.62	0.529	27	0.0142	0.361	0.102	174
3	0.229	5.83	26.67	0.667	28	0.0126	0.321	0.081	220
4	0.204	5.19	21.15	0.842	29	0.0113	0.286	0.0642	277
5	0.182	4.62	16.75	1.06	30	0.01	0.255	0.051	349
6	0.162	4.11	13.3	1.34	31	0.00893	0.227	0.0404	441
7	0.144	3.67	10.55	1.69	32	0.00795	0.202	0.032	556
8	0.128	3.26	8.37	2.13	33	0.00708	0.18	0.0254	701
9	0.114	2.91	6.63	2.68	34	0.00631	0.16	0.0201	884
10	0.102	2.59	5.26	3.38	35	0.00562	0.143	0.016	1114
11	0.0907	2.3	4.17	4.27	36	0.005	0.127	0.0127	1405
12	0.0808	2.05	3.31	5.38	37	0.00445	0.113	0.01	1772
13	0.072	1.83	2.62	6.78	38	0.00396	0.101	0.00797	2234
14	0.0641	1.63	2.08	8.55	39	0.00353	0.0897	0.00632	2818
15	0.0571	1.45	1.65	10.8	40	0.00315	0.0799	0.00501	3552
16	0.0508	1.29	1.31	13.6	41	0.0028	0.0711	0.00397	4481
17	0.0453	1.15	1.038	17.1	42	0.00249	0.0632	0.00314	5666
18	0.0403	1.024	0.823	21.6	43	0.00222	0.0564	0.0025	7128
19	0.0359	0.912	0.653	27.3	44	0.00197	0.05	0.00197	9052
20	0.032	0.812	0.518	34.4	45	0.00176	0.0447	0.00157	11,341
21	0.0285	0.723	0.41	43.4	46	0.00157	0.0399	0.00125	14,252
22	0.0254	0.644	0.326	54.7	47	0.0014	0.0355	0.00099	18,022
23	0.0226	0.573	0.258	68.9	48	0.00124	0.0316	0.00078	22,726
24	0.0201	0.511	0.205	86.9	49	0.00111	0.0281	0.00062	28,657
25	0.0179	0.455	0.162	110	50	0.00099	0.025	0.00049	36,137

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Appendix M
Rules of Thumb

Sometimes an OKAY answer NOW! is more important than a good answer LATE.

—Eric Bogatin

This chapter contains useful rules of thumb and simplified, easy-to-remember formulas for every EMC design engineer.

M.1 Air Breakdown Voltage

V ≈ 3 kV/mm
Figure M.1 shows that the breakdown voltage for air is around 5 kV/mm at sea level and 3 kV/mm at 5000 m altitude. To avoid unexpected voltage breakdowns, the air breakdown voltage is usually assumed to be 3 kV/mm.

M.2 Amplitude of Square Wave Harmonics

Assumption: ideal square wave with 0 sec rise-/fall-time and 50% pulse duration (see Fig. M.2).

$$V_{n,RMS} \approx \frac{4}{\pi} \cdot \frac{1}{n} \cdot \frac{V_{pp}}{2\sqrt{2}} \approx \frac{0.45 V_{pp}}{n} \tag{M.1}$$

where

$V_{n,RMS}$ = RMS value of the n-th harmonics sine wave in [V]
V_{pp} = peak-to-peak value of the ideal square wave in [V]

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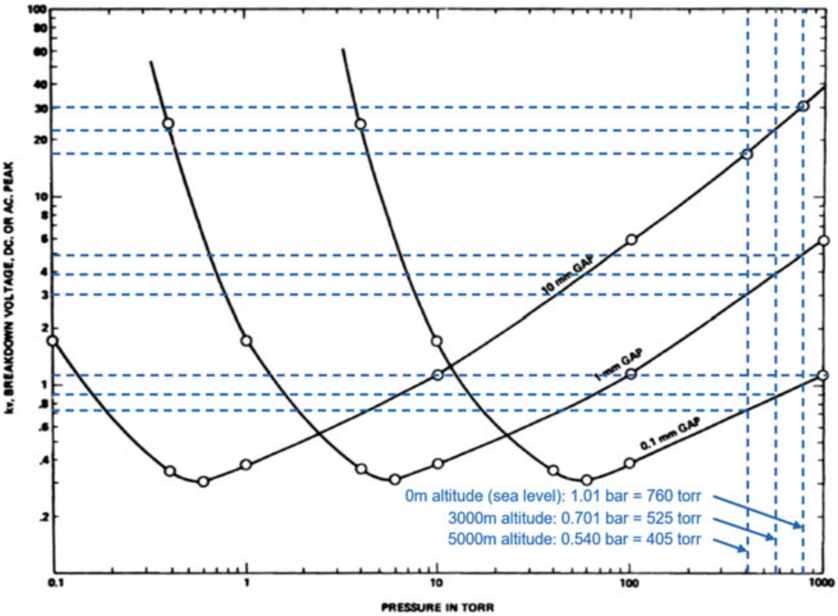

Fig. M.1 Paschen's original curves show the breakdown voltage of air as a function of pressure [1]

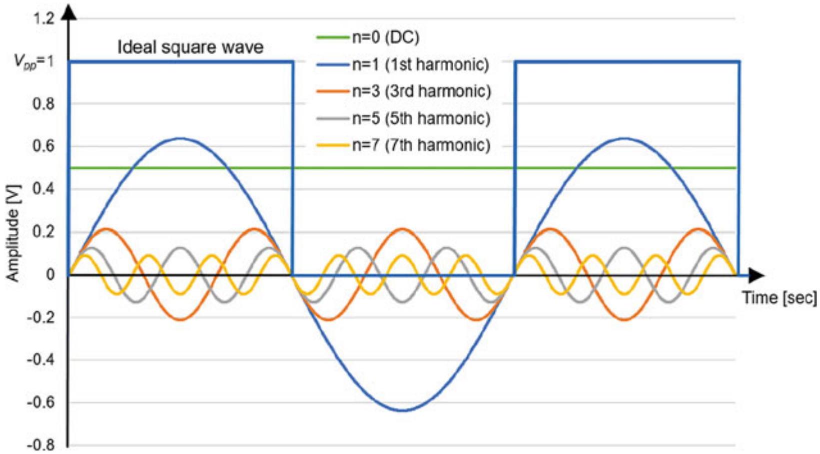

Fig. M.2 Ideal square wave with its first seven harmonics

n = odd harmonics number [1, 3, 5, ...], even harmonics have an amplitude of 0 V for an ideal square wave signal

M.3 Antenna Input Power for Desired E-Field Strength

$$P_t = \frac{(E \cdot d)^2}{30 \cdot G_{it}} = \frac{(E \cdot d)^2}{30 \cdot 10^{\frac{G_{it}[\mathrm{dBi}]}{10}}} \tag{M.2}$$

where

P_t = RMS input power at the transmitting antenna's input terminals for achieving electric field strength E at distance d (assuming free-space, line-of-sight, distance d in the far-field and in the direction of the transmitting antenna's main-lobe and matched polarization to the transmitting antenna) in [W]
E = RMS field strength of the electric field in [V/m]
G_{it} = antenna gain of the transmitting antenna compared to an isotropic radiator (dimensionless)
G_{it}[dBi] = antenna gain of the transmitting antenna compared to an isotropic radiator in [dBi]
d = distance from the transmitting antenna in [m]

M.4 Bandwidth of Digital Signals

$\mathbf{B} \approx \frac{\mathbf{0.35}}{\mathbf{t_{10\%-90\%}}}$
B [Hz] is the highest significant frequency in the spectrum of the digital signal and $t_{10\%-90\%}$ [sec] the minimum rise/fall-time of the digital signal from 10% to 90%.

M.5 Critical Length of Transmission Lines

$\mathbf{l_{critical}} > \lambda/10$
A transmission line, which is longer than a tenth of the signals wavelength λ, should no longer be modeled with lumped RLC-elements, but instead with its characteristic impedance Z_0 [Ω]. More details in Sect. 7.2.3 on page 66.

M.6 Characteristic Impedance of Wires

$\mathbf{Z_0} \approx 100\ \Omega$ to $150\ \Omega$
Typical characteristic impedance Z_0 [Ω] of a pair of wire is 100 Ω to 150 Ω [3].

M.7 Capacitance Per-Unit-Length

C' ≈ 50 fF/mm
Assumption: small wire (e.g., AWG26) or narrow PCB trace (e.g., 0.25 mm), isolated with plastics, forward and return path close to each other (< 1 mm). More details in Sect. 7.4 on page 73.

M.8 Inductance Per-Unit-Length

L' ≈ 1 nH/mm
Assumption: small wire (e.g., AWG26) or narrow PCB trace (e.g., 0.25 mm). The value represents the loop inductance per-unit-length: internal, external, and mutual inductances of forward and return current path. More details in Sect. 7.4 on page 73.

M.9 CMOS Input Impedance

$\mathbf{R_{in} > 1G\Omega}$
$\mathbf{C_{in} \approx 5pF}$
The input capacitance C_{in} [F] of integrated CMOS circuits is often used in combination with an external output resistor R_{ext} [Ω] close to the driver's output for controlling the rise- and fall-time of digital signals (see Fig. M.3).

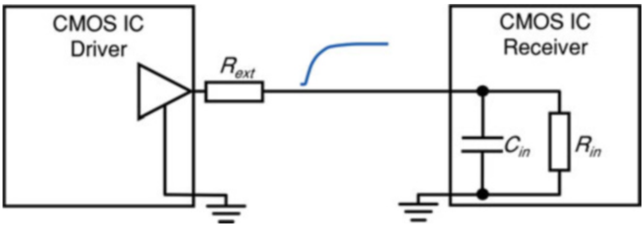

Fig. M.3 R_{out} [Ω] is used to control the voltage transients

M.10 E-Field from Differential-Mode Currents in Small Loops

$$E_{DM,max} \approx \begin{cases} 1.32 \cdot 10^{-14} \frac{I_{DM} f^2 A}{d} & \text{without reflecting ground plane} \\ 2.64 \cdot 10^{-14} \frac{I_{DM} f^2 A}{d} & \text{with reflecting ground plane (Fig. 9.7 on page 114)} \end{cases} \tag{M.3}$$

where

$E_{DM,max}$ = maximum RMS field-strength measured in the far-field and radiated by an electrically a small current loop with area A, circumference $< \lambda/4$ and driven by the differential current I_{DM} in [V/m] (Sect. 9.9.1 on page 121)
I_{DM} = RMS value of the differential-mode current in [A]
f = frequency of the sinusoidal current signal in [Hz]
A = area of the current loop in [m^2]
d = distance to the center of the current loop in [m]

M.11 E-Field from Common-Mode Currents in Short Cables

$$E_{CM,max} \approx \begin{cases} 0.63 \cdot 10^{-6} \frac{I_{CM} f l}{d} & \text{without reflecting ground plane} \\ 1.26 \cdot 10^{-6} \frac{I_{CM} f l}{d} & \text{with reflecting ground plane (Fig. 9.7 on page 114)} \end{cases} \tag{M.4}$$

where

$E_{CM,max}$ = maximum RMS field-strength measured in the far-field and radiated by an electrically short cable of length $l < \lambda/4$ and driven by the common-mode current I_{CM} in [V/m] (Sect. 9.9.2 on page 122)
I_{CM} = RMS value of the common-mode current along the cable in [A]
f = frequency of the sinusoidal current signal in [Hz]
l = length of the conductor or transmission line in [m]
d = distance to the center of the cable in [m]

M.12 E-Field from Common-Mode Current with Cable at Resonance

$$E_{max} \approx \begin{cases} 60 \frac{I_{CM}}{d} & \text{without reflecting ground plane} \\ 120 \frac{I_{CM}}{d} & \text{with reflecting ground plane (Fig. 9.7 on page 114)} \end{cases} \tag{M.5}$$

where

E_{max} = maximum RMS E-field at distance d for matched impedances, matched polarization, line-of-sight path, and a resonant structure (that is resonant at the frequency of I_{CM}) driven by I_{CM} in [V/m]
I_{CM} = RMS common-mode current that flows along the resonant structure in [A]
d = distance from the antenna, where E_{max} is measured in [m]

M.13 E-Field from Common-Mode Voltage with Cable at Resonance

$$E_{max} \approx \begin{cases} 0.8\frac{V_{CM}}{d} & \lambda/2\text{-dipole without reflecting ground plane} \\ 1.6\frac{V_{CM}}{d} & \lambda/2\text{-dipole with reflecting ground plane (Fig. 9.7 on page 114),} \\ 1.6\frac{V_{CM}}{d} & \lambda/4\text{-monopole without reflecting ground plane} \\ 3.2\frac{V_{CM}}{d} & \lambda/4\text{-monopole with reflecting ground plane (Fig. 9.7 on page 114)} \end{cases} \tag{M.6}$$

where

E_{max} = maximum RMS E-field at distance d for matched impedances, matched polarization, line-of-sight path, and a resonant structure (that is resonant at the frequency of V_{CM}) driven by V_{CM} in [V/m]
V_{CM} = RMS common-mode voltage that drives the resonant structure in [V]
d = distance from the antenna, where E_{max} is measured in [m]

M.14 E-Field Strength for Given Antenna Power

$$E = \frac{\sqrt{30 P_t G_{it}}}{d} = \frac{\sqrt{30 P_t 10^{\frac{G_{it}[\text{dBi}]}{10}}}}{d} \tag{M.7}$$

where

E = RMS field strength of the electric field at distance d from the transmitting antenna for input power P_t (assuming free-space, line-of-sight, distance d in the far-field and in the direction of the transmitting antenna's main-lobe and matched polarization to the transmitting antenna) in [V/m]
P_t = RMS power input to the transmitting antenna in [W]
G_{it} = antenna gain of the transmitting antenna compared to an isotropic radiator (dimensionless)
G_{it}[dBi] = antenna gain of the transmitting antenna compared to an isotropic radiator in [dBi]

d = distance from the transmitting antenna in [m]

M.15 Galvanic Corrosion

To prevent galvanic corrosion of the connection of dissimilar metals, the galvanic series voltage difference of the dissimilar metals should be $\leq 0.5\,\mathrm{V}$ for controlled environments and $\leq 0.15\,\mathrm{V}$ for harsh environments [2]. More details in Chap. H on page 311.

M.16 Return Current on PCB Planes

The return current path on PCBs with a solid reference plane [3]:

- **f < 100 kHz:** return current spreads out over the whole reference plane.
- **f > 1 MHz:** return current flows directly underneath the forward signal trace.

M.17 Via Capacitance and Inductance

$\mathbf{C_{via} \approx 0.4\ pF}$[1]
$\mathbf{L_{via} \approx 1.2\ nH}$[2]

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[1] $C_{via} \approx \frac{0.0555\varepsilon_r' h d_1}{d_2 - d_1}$, where h = thickness of PCB in [mm], d_1 = diameter of the via pad in [mm], d_2 = diameter of the anti-pad (void area between the pad and the copper plane) in [mm] and C_{via} is the approximate via capacitance in [pF] [4].

[2] $L_{via} \approx \frac{h}{5} \cdot \left(1 + \ln\left(\frac{4h}{d}\right)\right)$, where h = thickness of PCB in [mm], d = inner diameter of via in [mm] and L_{via} is the approximate via inductance in [nH] [4].

Index

Symbols
1 dB compression point (P1dB), 56

A
Abbreviations, xxiii
Acronyms, xxiii
American Wire Gauge (AWG), 401
Ampére's law, 301
Analog-To-Digital Converter (ADC), 259
ANATEL label, 326
Antenna factor, 116
Antennas, 111
  $\lambda/2$-dipoles, 114
  $\lambda/4$-monoples, 114
  antenna gain, 113
  coil factor, 123
  directivity, 113
  E-field antennas, 111, 119, 120
  effective aperture, 116
  EIRP, 114
  ERP, 114
  far-field antennas, 112
  H-field antenna, 122
  induction coil, 123
  intended antennas, 119
  isotropic radiator, 112
  maximum effective aperture, 116
  near-field antennas, 111
  realized antenna gain, 115
  unintended antennas, 125
Attenuation constant, 83

B
Balanced transmission line, 89
Balun, 90, 178
Bandwidth, 45
Bandwidth of digital signals, 45, 405
Bode plot, 248
Breakdown voltage air, 403
BSMI mark, 334
Burst filters, 256
Bursts, 377

C
Cable, 37
Cable shields, 225
Capacitance of vias, 409
Capacitance per-unit-length, 406
Capacitive coupling, 192, 202, 204
Capacitors, 157, 258
  aging, 163
  decoupling, 157
  frequency dependence, 163
  multilayer ceramic capacitors, 157
  safety capacitors, 158
  temperature derating, 160
  voltage derating, 161
Causal system, 245
CB scheme, 21
CCC mark, 327
CE mark, 329
CH mark, 333
Characteristic impedance, 69, 85
Characteristic impedance of wires, 405

R. B. Keller, *Design for Electromagnetic Compatibility–In a Nutshell*,
https://doi.org/10.1007/978-3-031-14186-7

CISPR 11, 344
CISPR 32, 357
Clamping devices, 180
Common impedance coupling, 191, 201, 236, 266
Common-mode, 198
Common-mode chokes, 173
Common-mode current, 200, 203, 206
Common-mode noise, 203
Common-mode to differential-mode conversion, 203
Common-mode voltage, 199, 206
Complex conjugate, 281
Complex numbers, 279
Complex permeability, 290
Complex permittivity, 289
Compliance marks, 19
Compliance testing, 344
Conducted immunity, 4, 388
Conducted immunity testing, 384
Conducted susceptibility, 4
Conductive coupling, 191
Conductivity, 307
Conductor, 145, 258
Conductor equivalent circuits, 145
Conformity assessment, 9
Constants, 277
Coplanar wave guide, 38
Corrosion, 315
Coupling
    capacitive coupling, 202
    conductive coupling, 201
    electromagnetic coupling, 197
    inductive coupling, 195
Coupling path
    capacitive coupling, 192
Coupting paths, 189
Critical transmission line length, 405
Crowbar devices, 185
Curie temperature, 166
Current, 288
Current density, 294

D
Date Of Withdrawal (DOW), 331
Decibel, 24
Decoupling, 157, 267
    global decoupling, 267
    local decoupling, 267
Design guidelines, 265
Dielectric constant, 277, 289, 310
Differential-mode, 198
Differential-mode noise, 200
Differential-mode to common-mode conversion, 205
Differential-mode voltage, 199
Dirac delta function, 246
Directivity, 113
Displacement current, 293
Displacement current density, 293
Dissipation factor, 88
Distributed parameter model, 69

E
EAC mark, 328
Effective aperture, 116
Electric current density, 294
Electric displacement field, 292
Electric field, 291
Electric field coupling, 192
Electric flux, 292
Electric flux density, 292
Electrical Fast Transients (EFT), 256, 377
Electromagnetic Compatibility (EMC), 1
    emission, 2
    immunity, 2
    susceptibility, 2
Electromagnetic coupling, 197, 205
Electromagnetic field, 95, 297
Electromagnetic Interference (EMI), 1
Electromagnetic Pulse (EMP), 4, 115
Electromagnetism, 287
Electromotive Force (EMF), 300
Electrostatic Discharge (ESD), 4, 321, 369
EMC compliance, 10
EMC design guidelines, 265
EMC Directive (EMCD), 10, 329
EMC regulations, 10, 325
EMC standards, 11
    basic EMC publications, 16
    EMC emission standards, 14
    EMC immunity standards, 14
    EMC product family standards, 17
    EMC product standards, 17
    generic EMC standards, 17
    performance criteria, 15
EMC test lab, 21
EMC testing, 343
    CB EMC test scheme, 21
    emission testing, 33
    immunity testing, 33
EMI receiver, 344
Emission
    conducted emission, 3
    radiated emission, 3

Emission testing, 119, 122, 344, 357
Equivalent Series Inductance (ESL), 157
Equivalent Series Resistance (ESR), 157
ESD filters, 256
ETSI, 13, 330
Euclidean geometry, 283
Euler's number, 277
Eurasian economic union 13g, 328
European Committee for Standardization (CEN), 330
European Committee for Electrotechnical Standardization (CENELEC), 13, 330
European Telecommunications Standards Institute (ETSI), 13, 329, 330
European Union (EU), 329

F
Far-field, 100, 118, 214, 216
  near-field to far-field boundary, 98
Far-field coupling, 197
Faraday's law, 300
Fast Fourier Transform (FFT), 304
Federal Communications Commission (FCC), 336
Ferrite beads, 169
Ferrites, 169, 258
  cable mount ferrite beads, 169
  chip ferrite beads, 171
  ferrite beads, 169
Ferromagnetic materials, 307
Filter characterization, 245
Filtering, 245
Filters, 245
  active filters, 251
  analog filters, 259
  anti-aliasing filter, 259
  band-pass filters, 250
  common-mode filters, 254
  differential-mode filters, 254
  digital filters, 259
  FIR filters, 259
  high-pass filters, 250
  low-pass filters, 249
  passive filters, 251
  transient suppression filters, 256
Flicker, 3, 365
Fourier analysis, 41
Fourier series, 42
Fourier transform, 42, 248, 304, 403
  DFT, 42, 304
  discrete fourier transform, 304
  discrete-time fourier transform, 305
  DTFT, 305
  FFT, 43, 304
  fourier series, 303
FR-4, 312
Fraunhofer region, 100
Free-Space Path Loss (FSPL), 131
Frequency, 34
Frequency response, 245, 248
Frequency-domain, 41
Fresnel zone, 97
Friis transmission equation, 131

G
Galvanic corrosion, 409
Galvanic series, 317
Gas Discharge Tubes (GDT), 185
Gas discharge tubes (GDT), 258
Gauss' law for electricity, 299
Gauss' law for magnetism, 300
Ground bounce, 271
Ground loops, 241
Ground planes, 237
Ground system categories, 238
Grounding, 235
  chassis ground, 235
  daisy-chain ground system, 239
  functional ground, 235
  ground symbols, 235
  grounding of cable shields, 231
  grounding of shielding enclosures, 224
  hybrid cable shield grounding, 231
  safety ground, 235
  signal ground, 235
  single-point ground systems, 239
  star ground system, 239

H
Harmonic currents, 358
Harmonics amplitude, 403
Harmonised standards, 330
Hybrid cable grounding, 273
Hybrid ground systems, 240

I
IEC 61000-3-2, 358
IEC 61000-3-3, 365
IEC 61000-4-11, 392
IEC 61000-4-2, 369
IEC 61000-4-3, 373
IEC 61000-4-4, 377
IEC 61000-4-5, 380
IEC 61000-4-6, 384

IEC 61000-4-8, 388
Imbalance Difference Modeling (IDM), 205
Imbalance factor, 199, 200, 206
Immunity testing, 120, 123
Impedance, 288
Impedance matching, 61
  s_{11}, 50
  reflection coefficient, 49
  return loss, 51
  S-parameters, 53
  VSWR, 50
Impulse response, 245, 246
Indian Standards Institution (ISI) mark, 331
Inductance
  external inductance, 152
  internal self-inductance, 149
  loop inductance, 154
  mutual inductance, 147
  partical inductance, 153
Inductance of vias, 409
Inductance per-unit-length, 406
Inductive coupling, 195, 202
Inductors, 164
  Curie temperature, 166
  current saturation, 166
  frequency dependence, 166
Innovation, Science and Economic Development (ISED), 327
Insertion loss, 52
Insulators, 310
Integrad transform, 303
Intended antenna, 119
Internal impedance, 151
  internal inductance, 149
  resistance, 148
International Electrotechnical Commission (IEC), 11
International Electrotechnical Commission for Electrical Equipment (IECEE), 21
International Telecommunication Union (ITU), 34
  frequency bands, 34
  frequency spectrum, 34
Intrinsic impedance, 100
Intrinsic impedance of vacuum, 277
In-Vitro Diagnostics Regulation (IVDR), 329
ISED certification number, 327
ISM frequency bands, 345
Isotropic radiator, 112

K
Korean Certification (KC) mark, 333

L
Laplace transform, 43, 248, 305
Life cycle costs, 5
Linear and Time-Invariant (LTI) system, 245
Linear system, 245
Link budget, 132
Lorentz force, 302
Loss tangent, 88

M
Magnetic field, 295
Magnetic field coupling, 195
Magnetic flux, 295
Magnetic flux density, 296
Magnetic induction, 296
Magnetic permeability, 277, 307
Material properties, 307
Maxwell's equations, 299
Medical Device Regulation (MDR), 329
Microstrip line, 38
Mode conversion, 203, 205
Multilayer Ceramic Capacitors (MLCCs), 157
Multipath effect, 119
Multi-point ground systems, 239
Mutual inductance, 147

N
Near-field, 97, 216
  near-field to far-field boundary, 98
  radiating near-field, 99
  reactive near-field, 99
Near-field coupling, 192, 196
Noise coupling, 190
Noise factor, 55
Noise figure, 55
Nyquist-Shannon sampling theorem, 259

O
One-port circuit, 50

P
PCB laminates, 312
PCB materials, 311
Permeability, 290, 307
Permeability of vacuum, 277
Permittivity, 289, 310
Permittivity of vacuum, 277
Phase constant, 83
Polarization, 294
Polarization current, 293
Polarization density, 294

Power, 289
Power density, 112, 297
Poynting vector, 95
Pre-compliance testing, 343
Presumption of conformity, 331
Printed Circuit Board (PCB), 37, 311
  PCB trace, 37
Propagation constant, 82
Proximity effect, 153
PSE label, 332

R
Radiated coupling, 197
Radiated emissions
  common-mode signals, 130
  maximum e-field from common-mode signals, 127
  short cables, 126
  small circuit loops, 125
Radiated immunity, 4
Radiated immunity testing, 373
Radiated susceptibility, 4
Radio Equipment Directive (RED), 329
RCM mark, 325
Reactance, 288
Reflection coefficient, 49
Reflection parameter, 61
Regulatory Compliance Mark (RCM), 325
Resistance, 148, 288
Resistivity, 307
Resistors, 154, 258
Resolution bandwidth, 58
Return current path, 237
Return loss, 51
Ringing, 46
Rules of thumb, 403

S
Safety, 21
Sampling frequency, 259
Scattering parameters, 53
Self-inductance, 165
Sheet resistance, 140
Shielding, 212
  absorption loss, 213
  cable shields, 225
  grounding of cable shields, 231
  grounding of shielding enclosures, 224
  hybrid cable shield grounding, 231
  low-frequency magnetic field shielding, 219
  multiple-reflection loss, 213
  reflection loss, 213
  slots and apertures, 219
  transfer impedance, 225
  waveguide below cutoff, 223
Shielding effectiveness, 212
Signal-To-Noise Ratio (SNR), 55
Skin depth, 135
S-parameters, 53
Specific conductance, 307
Spectrum analyzers, 57, 344
  average detector, 61
  center frequency, 58
  EMI receivers, 58
  peak detector, 61
  quasi-peak detector, 61
  resolution bandwidth, 58
  RMS detector, 61
  span, 58
  start frequency, 58
  stop frequency, 58
  sweep time, 58
  video bandwidth, 58
Speed of light, 277
SPICE simulation, 251
Standard electrode potentials, 316
Step response, 247
Stitching vias, 271
Surface resistance, 140
Surge, 380
Surge filters, 256
Symbols, xxvii

T
Technical Committee 77 (TC 77), 11
Thyristors, 185, 258
Time-domain, 41
Time-invariant system, 245
Transfer impedance, 225
Transient filters
  capacitors, 258
  clamping devices, 180
  crowbar devices, 185
  gas discharge tubes, 185, 258
  TVS diodes, 183, 258
  TVS thyristors, 185, 258
  varistors, 180, 258
Transient suppression filters, 256
Transient-Voltage-Suppression Diodes (TVS diodes), 183, 258
Transmission line input impedance, 85

Transmission lines, 38, 65
  attenuation constant, 83
  balanced line, 66
  balanced transmission line, 89, 90
  characteristic impedance, 69, 85
  coaxial cable, 65
  critical length, 68
  dielectric losses, 87
  differential signal interfaces, 90
  differential signaling interfaces, 90
  dissipation factor, 88
  distributed parameter model, 69
  high-frequency losses, 87
  loss tangent, 88
  lossless transmission line, 70, 85
  microstrip line, 65
  ohmic losses, 87
  per-unit-length capacitance, 75
  per-unit-length inductance, 75
  phase constant, 83
  propagation constant, 82
  single-ended interfaces, 90
  stripline, 65
  unbalanced transmission line, 89, 90
  waveguide, 66
Triboelectric effect, 321
Turkish Standards (TSE) mark, 335
Twisted pair, 38
Two-port circuit, 53

U
Unbalanced transmission line, 89
The United Kingdom Conformity mark (UKCA mark), 336
United Nations Economic Commission for Europe Regulation 10 (UNECE R 10), 329
Units, xxvii
Unit step, 247

V
Varistors, 180, 258
Vector geometry, 283
  cross product, 285
  dot product, 284
Vector network analyzer, 54
Velocity factor, 38
Via
  capacitance, 409
  inductance, 409
Video bandwidth, 58
Voltage, 287
Voltagedips, 392
Voltage-Dependent Resistors (VDR), 180, 258
Voltage Standing Wave Ratio (VSWR), 50, 61

W
Wave impedance, 102
  far-field, 106
  near-field, 104, 105
Waveguide, 66, 223
Wavelength, 34, 37

Z
Z-transform, 43, 306