SCHOOL
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Faculty of Science and Engineering

MACQUARIE University

APEC 2018 Professional Education Seminar

Electromagnetic Interference & Compatibility for Power Electronics Engineers

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OVERVIEW

- **INTRODUCTION - speaker's background**
- **OVERVIEW OF TUTORIAL**
  - **PART 1 – Context**
    - Electromagnetic interference & compatibility
    - Trends in power systems and power electronics
    - Potential impacts of EMI in wireless communication
    - EMI Standards
  - **PART 2 – Fundamentals of EMI**
    - Power electronics as source of EMI
    - Spectral characteristics of switching signals
    - Electromagnetic coupling (conduction, induction, radiation)
  - **PART 3 - Principles and practice of EMI minimisation**
    - Minimising EMI at the source (e.g. PWM and switching techniques)
    - Minimising EMI coupling (e.g. circuit layout, filtering, shielding)
    - Measuring EMI

About the presenter…

Graham Town

- Professor, School of Engineering, Macquarie University
- graham.town@mq.edu.au

Presenter Overview

Graham Town is an electrical engineer with 8 years experience in the Australian electronics industry, and 30 years experience in engineering education and research. He received a Bachelor of Electrical Engineering (Hons1) from NSWIT (now UTS) in 1984, a PhD in medical imaging from the University of Sydney in 1992, and a Graduate Certificate in Leadership and Management (Higher Education) from Macquarie University in 2007.

Graham is currently a Professor in the School of Engineering at Macquarie University where he established Macquarie University's undergraduate engineering program, first offered in 2004. He has published extensively in diverse areas including medical imaging, terahertz technology, guided-wave optics and photonics, and in recent years has been leading industry-supported research on power electronics, and smart-grids with a focus on electric vehicles.

Dr Town is a Senior Member of the IEEE, and a Fellow of Engineers Australia.

About the presenter

- **1978-81/83/85: Radio Trades Apprentice / Engg Trainee / Engineer**
  - **AWA Pty Ltd, Industrial Products Division**
  - **Interscan, μwave comm’s, 1st generation optical fibre comm’s, etc.**
- **1978-84: BE (Hons 1) NSWIT in Electrical Engineering**
  - **Thesis on holographic antenna measurements, CSIRO Radiophysics**
- **1985 – 1991: PhD Sydney Univ. in Medical Imaging**
  - **Built 1.5T (64MHz) NMR imaging system**
- **1992 - 2002: Academic Engineer, Sydney Univ.**
  - **3rd gen’n optical fibre lasers & guided-wave devices for telecomms and sensing**
- **2002 - : Academic Engineer, Macquarie Univ.**
  - **Established BE program at Macquarie Univ. (commenced 2004)**
  - **Member Sustainable Energy Systems Engineering research group**
  - **Some externally funded research projects since 2010…**
    - **Integrated Energy Conversion (GaN power electronics)**
    - **Distributed Energy Storage and Management (EVs in electrical energy systems)**

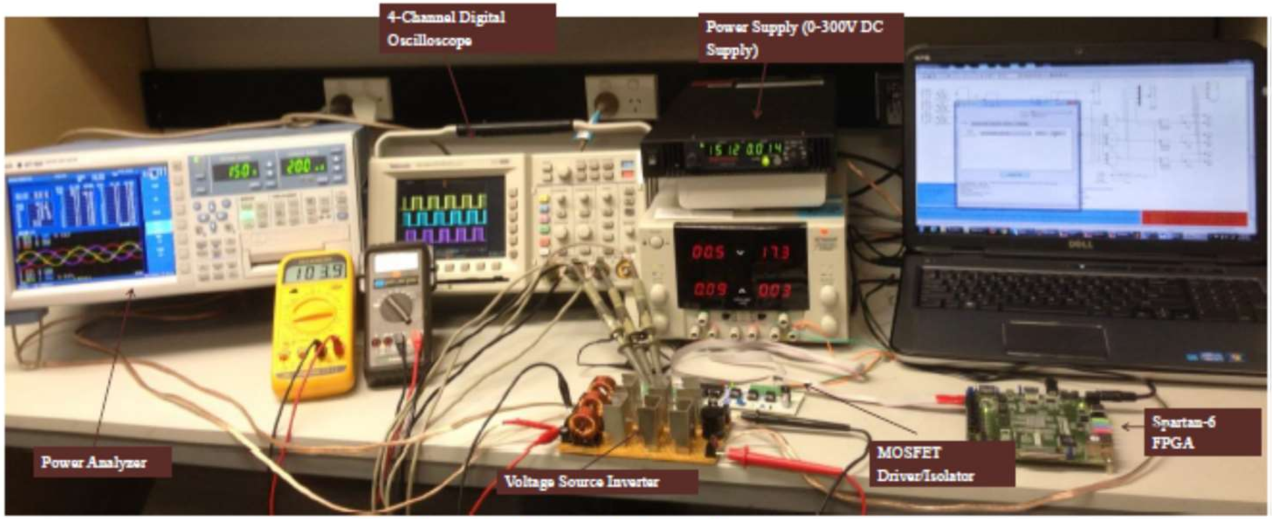
Integrated Energy Conversion

Motivations: Trend to increasingly compact and portable electronic devices. Increases in energy efficiency → savings in infrastructure.

Goal: Electronic power converters with increased efficiency, decreased size.

Methods: GaN devices and circuits, RF switching, novel circuit topologies, intelligent control (e.g. efficient PWM, EMI minimisation, etc).

Applications: Distributed power systems & microgrids (e.g. IT server farms), renewable energy systems, energy-efficient lighting, electric vehicles.



Testing FPGA-controlled switch-mode power supply

Distributed Energy Storage

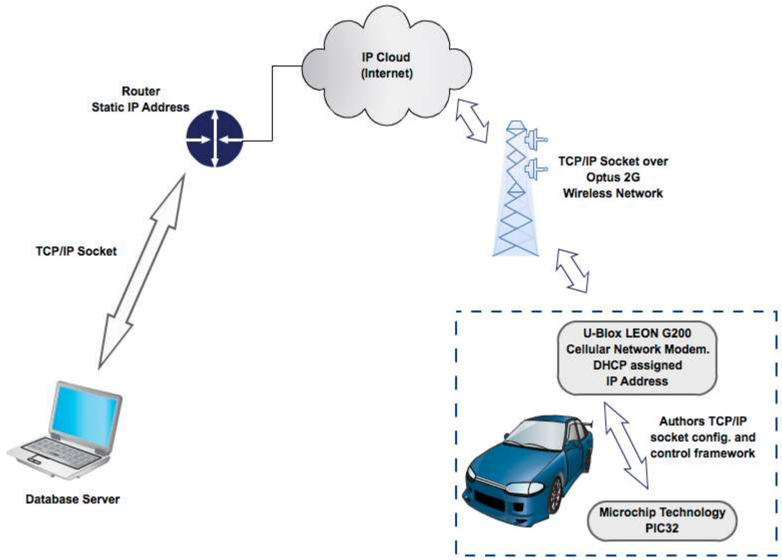
Motivations:

~ 35% of energy in developed countries used for transport
~ 20% of vehicles in Australia to be fully or hybrid electric by 2020 [AECOM2012].

Goal: Management electric vehicles and their impact on electric power grids.

Methods: Automatic (M2M) wireless monitoring of vehicle state of charge and position, "smart charging" algorithms for EV batteries, new services for EVs and/or enabled by EVs.

Applications: Future "smart-grid" energy systems incorporating renewable energy and electric vehicles.



PART 1 - Overview - Aims of this tutorial

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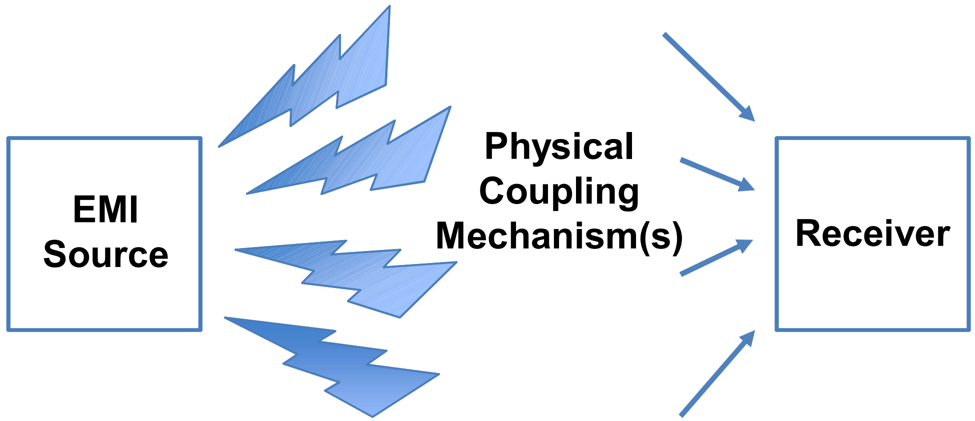
- **Assist electrical engineers understand → apply**
  - **PART 1 - Characteristics of EMI**
    - how/why EMI can occur, especially in wireless communication systems
    - potential signs and effects of EMI
    - standards regarding limits and measurements of EMI
  - **PART 2 - Physical basis of EMI in power electronics**
    - switching signals (time ←→ frequency)
    - electromagenetics (circuits ←→ EM fields)
    - EMI coupling mechanisms (conducted, inducted, radiated) and their dependence upon frequency, distance
  - **PART 3 - Methods for measuring and minimising EMI**
    - by reducing EMI at source
    - by impeding coupling of EMI

EMI and EMC - Definitions



- **Electromagnetic Interference (EMI)**

"Any electromagnetic disturbance, induced intentionally or unintentionally, that interrupts, obstructs, or otherwise degrades or limits the effective performance of electronics and electrical equipment." *Dictionary of Military and Associated Terms. US Department of Defense 2017.*

The effect of unwanted energy due to one or a combination of emissions, radiations, or inductions upon reception ...., manifested by any performance degradation, misinterpretation, or loss of information which could be extracted in the absence of such unwanted energy" *(ITU Radio Regulations, Section IV. Radio Stations and Systems – Article 1.166)*

See also https://en.wikipedia.org/wiki/Electromagnetic_interference

- **Electromagnetic compatibility (EMC)**

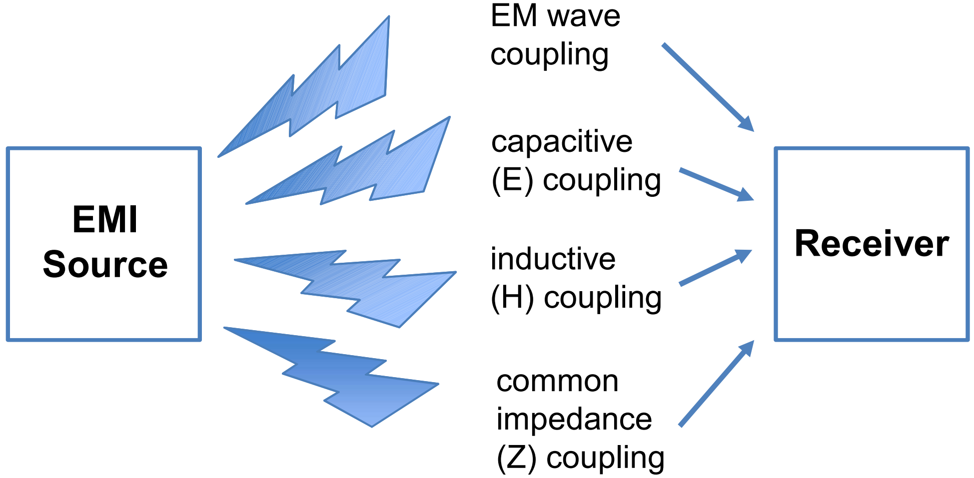
"The ability of systems, equipment, and devices that use the electromagnetic spectrum to operate in their intended environments without causing or suffering unacceptable or unintentional degradation because of electromagnetic radiation or response." *Dictionary of Military and Associated Terms. US Department of Defense 2017.*

See also https://en.wikipedia.org/wiki/Electromagnetic_interference

EMI and EMC
- Overview

- **Reducing EMI usually improves EMC (reciprocity)**
- **Options for limiting EMI in any system**
  1. **Reduce interfering emissions from source**
  2. **Impede coupling(s) of interfering emissions**
  3. ~~**Reduce susceptibility of receiver to interfering emissions**~~

Lightning
Radio transmitters
Electrical equipment
(e.g. motors, lighting)
Computers
Microwave oven
Power electronics



Biological organisms
Wireless receivers
Electronic equipment
Sensors (many applications)

➢ The variety of source characteristics and coupling mechanisms can make EMI difficult to diagnose - the effects are often intermittent and remote from the source - but the impacts are potentially serious.

CONTEXT
- Types and sources of EMI

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- **Common types and typical sources of EMI…**
  - Electrostatic discharge (transient)
    e.g. lightning, failing HV electrical infrastructure, etc.
  - Voltage fluctuations and flicker (transient)
    e.g. sudden changes in large and reactive loads, e.g. motors.
  - Broadband radio-frequency emissions caused by electrical switching power electronics, motors, etc.
    e.g. “hard” switching in power electronics
- **Note: impacts, characterisation methods, standards, and mitigation techniques usually different for different types of EMI**

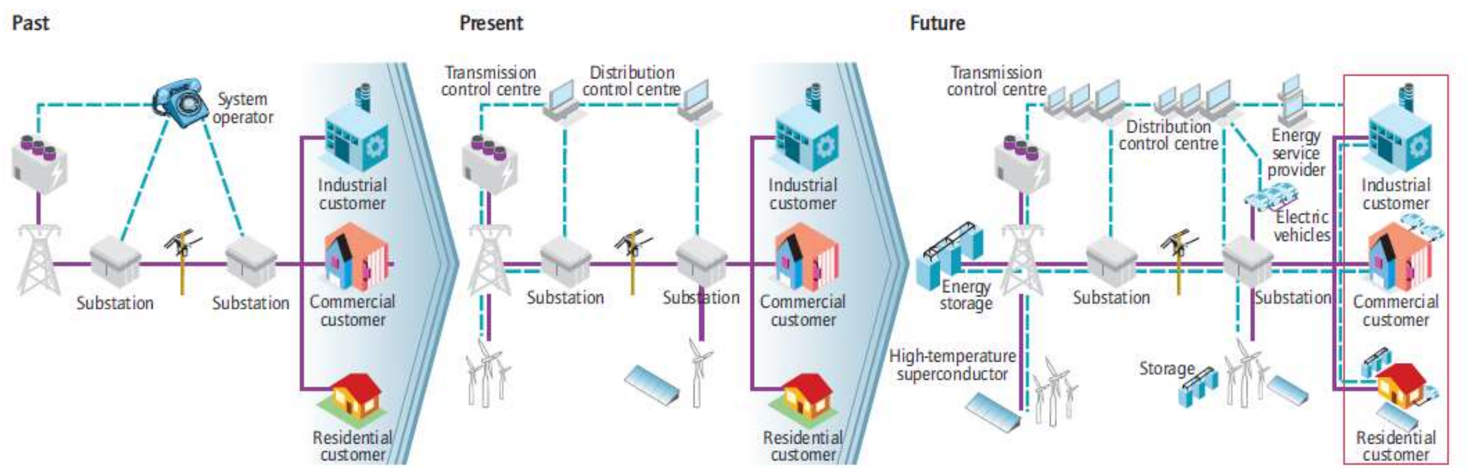
CONTEXT
- FOCUS, TRENDS

- **This tutorial will focus on EMI…**
  - **caused by power electronics**
  - **in electronic and wireless communication systems…**

- **Motivations…**
  - **trends to increasingly compact and/or efficient power electronics**
    - higher switching frequencies, higher speed switching devices (e.g. GaN)
  - **trends to increasingly pervasive power electronics**
    - increased efficiency, control of electrical equipment, e.g. solid state lighting, motors, etc.
  - **trends to increasing importance of wireless communications in the Internet of Things (IoT) for applications in monitoring and control**
    - "smart" grids, microgrids, intelligent transport, etc.
    - emerging low power wide area network standards, use unlicensed (ISM) RF bands, etc.
  - **trends in EMI regulation and control**
    - onus shifting to suppliers to ensure compliance with emission limits.
  - **costs associated with EMI/EMC minimized if dealt with at the design stage**

Background: McHenry, Roberson and Matheson, IEEE Spectrum, August 2015. http://spectrum.ieee.org/telecom/wireless/electronic-noise-is-drowning-out-the-internet-of-things

CONTEXT
- Trends in Power Systems



Smart Grids: Technology Roadmap, IEA 2011.

Characteristics and advantages of smart grids

Smart appliances	**Smart operations**
• Controllable loads • Status reporting • EV as sink/source (V2G)	• Ancillary services • Reliability • Service connections • Firming of renewables
Smart pricing	**Smart Planning**
• Value differentiated pricing • Aggregation pricing • Market participation	• Deferral of infrastructure • Minimise cost • Minimise carbon

CONTEXT
- Trends in Power Electronics

Widebandgap devices - high switching rate and frequency

→ increased switching efficiency

→ increased power density (smaller reactive components)

- GaN HEMTS can switch very fast (observed):
  - 600 V / 2 ns = 300 kV/us
  - 200 A / 2 ns = 100 kA/us
- Can in turn cause large transient in i & v through parasitic elements:
  - $i_C = C. dv/dt$ if $C = 10$ pF, $i_{Cpk} = 3$ A
  - $v_L = L. di/dt$ if $L = 1$ nH, $v_{Lpk} = 100$ V

→ **More efficient switching, but more EMI…**

- bandwidth (larger d/dt → 1000s harmonics, into GHz range)
- transient amplitude (increased peak i_C, v_L in parasitics; C, L)

CONTEXT
- Standards, Regulations

Standards and regulations are evolving with technology, policy, etc.

- **Power supply efficiency…**
  - e.g. US DOE Energy Conservation Standards for External Power Supplies, 2014.
    minimum efficiency typ. > 85%, requires switchmode supplies
- **EMI/EMC standards, limits on EMI generation…**
  - e.g. European Electromagnetic Compatibility (EMC) Directive 2014/30/EU
  - e.g. IEC 61000-3, CISPR 11 – 15
- **EMI/EMC measurement/characterisation methods…**
  - e.g. IEC 61000-4, CISPR 16, MIL-STD-461
- **Wireless communication systems…**
  - e.g. narrowband low power techniques (LTE-NB, etc.)
  - e.g. use of unlicensed (ISM) bands in RF spectrum

Note: See References section for links

CONTEXT
- Standards, Regulations

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ANSI		
	C63.4	Methods of Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz
IEC		
	IEC 60050-161	International Electrotechnical Vocabulary. Chapter 161: Electromagnetic compatibility
	IEC 60601-1-2	Medical electrical equipment - Part 1-2: General requirements for basic safety and essential performance - Collateral Standard: Electromagnetic disturbances - Requirements and tests
	IEC 60870-2-1	Telecontrol equipment and systems - Part 2: Operating conditions - Section 1: Power supply and electromagnetic compatibility
	IEC 60940	Guidance information on the application of capacitors, resistors, inductors and complete filter units for electromagnetic interference suppression
	IEC/TR 61000-1-1	Electromagnetic compatibility (EMC) - Part 1: General - Section 1: Application and interpretation of fundamental definitions and terms
	IEC/TS 61000-1-2	Electromagnetic compatibility (EMC) - Part 1-2: General - Methodology for the achievement of the functional safety of electrical and electronic equipment with regard to electromagnetic phenomena IEC/TR 61000-1-6 Electromagnetic compatibility
	IEC/TR 61000-2-1	Electromagnetic compatibility (EMC) - Part 2: Environment - Section 1: Description of the environment - Electromagnetic environment for low-frequency conducted disturbances and signaling in public power supply systems
	IEC 61000-3-8	Electromagnetic compatibility (EMC) - Part 3: Limits - Section 8: Signaling on low-voltage electrical installations - Emission levels, frequency bands and electromagnetic disturbance levels
	IEC/TR 61000-3-15	Electromagnetic compatibility (EMC) - Part 3-15: Limits - Assessment of low frequency electromagnetic immunity and emission requirements for dispersed generation systems in LV network
	IEC TR 61000-4-1	Electromagnetic compatibility (EMC) - Part 4-1: Testing and measurement techniques - Overview of IEC 61000-4 series
	IEC 61000-4-3	Electromagnetic compatibility (EMC) - Part 4-3 : Testing and measurement techniques - Radiated, radio-frequency, electromagnetic field immunity test
	IEC 61000-4-6	Electromagnetic compatibility (EMC) - Part 4-6: Testing and measurement techniques - Immunity to conducted disturbances, induced by radio-frequency fields
	IEC 61000-4-16	Electromagnetic compatibility (EMC) - Part 4-16: Testing and measurement techniques - Test for immunity to conducted, common mode disturbances in the frequency range 0 Hz to 150 kHz
	IEC 61000-4-19	Electromagnetic compatibility (EMC) - Part 4-19: Testing and measurement techniques - Test for immunity to conducted, differential mode disturbances and signalling in the frequency range 2 kHz to 150 kHz at a.c. power ports
	IEC/TR 61000-5-1	Electromagnetic compatibility (EMC) - Part 5: Installation and mitigation guidelines - Section 1: General considerations - Basic EMC publication
	IEC/TR 61000-5-2	Electromagnetic compatibility (EMC) - Part 5: Installation and mitigation guidelines - Section 2: Earthing and cabling
	IEC 61000-6-3	Electromagnetic compatibility (EMC) - Part 6-3: Generic standards - Emission standard for residential, commercial and light-industrial environments
	IEC 61000-6-4	Electromagnetic compatibility (EMC) - Part 6-4: Generic standards - Emission standard for industrial environments
	IEC 61326-1	Electrical equipment for measurement, control and laboratory use – EMC requirements – Part 1: General requirements
	IEC 61326-2-1	Electrical equipment for measurement, control and laboratory use - EMC requirements - Part 2-1: Particular requirements - Test configurations, operational conditions and performance criteria for sensitive test and measurement equipment for EMC unprotected applications
	IEC 61800-3	Adjustable speed electrical power drive systems - Part 3: EMC requirements and specific test methods
	IEC 62040-2	Uninterruptible power systems (UPS) - Part 2: Electromagnetic compatibility EMC) requirements
	IEC 62041	Power transformers, power supply units, reactors and similar products - EMC requirements

CISPR		
	CISPR 11	Industrial, scientific and medical (ISM) radio-frequency equipment - Electromagnetic disturbance characteristics - Limits and methods of measurement
	CISPR 14-1	Electromagnetic compatibility - Requirements for household appliances, electric tools and similar apparatus - Part 1: Emission
	CISPR 15	Limits and methods of measurement of radio disturbance characteristics of electrical lighting and similar equipment
	CISPR 16-1-1	Specification for radio disturbance and immunity measuring apparatus and methods - Part 1-1: Radio disturbance and immunity measuring apparatus - Measuring apparatus
	CISPR 16-1-2	Specification for radio disturbance and immunity measuring apparatus and methods - Part 1-2: Radio disturbance and immunity measuring apparatus - Coupling devices for conducted disturbance measurements
	CISPR 16-2-1	Specification for radio disturbance and immunity measuring apparatus and methods - Part 2-1: Methods of measurement of disturbances and immunity - Conducted disturbance measurements
	CISPR 16-2-3	Specification for radio disturbance and immunity measuring apparatus and methods - Part 2-3: Methods of measurement of disturbances and immunity - Radiated disturbance measurements
	CISPR 17	Methods of measurement of the suppression characteristics of passive EMC filtering devices
MIL		
	MIL-STD-461	REQUIREMENTS FOR THE CONTROL OF ELECTROMAGNETIC INTERFERENCE CHARACTERISTICS OF SUBSYSTEMS AND EQUIPMENT
	CE101	Conducted Emissions, Power Leads. 30 Hz to 10 kHz
	CE102	Conducted Emissions, Power Leads, 10 kHz to 10 MHz
	CE106	Conducted Emissions, Antenna Terminal, 10 kHz to 40 GHz
	CS101	Conducted Susceptibility, Power Leads, 30 Hz to 150 kHz
	CS103	Conducted Susceptibility, Antenna Port, Intermodulation, 15 kHz to 10 GHz
	CS104	Conducted Susceptibility, Antenna Port, Rejection or Undesired Signals, 30 Hz to 20 GHz
	CS105	Conducted Susceptibility, Antenna Port, Cross-Modulation, 30 Hz to 20 GHz
	CS109	Conducted Susceptibility, Structure Current, 60 Hz to 100 kHz
	CS114	Conducted Susceptibility, Bulk Cable Injection, 10 kHz to 200 MHz
	CS115	Conducted Susceptibility, Bulk Cable Injection, Impulse Excitation
	CS116	Conducted Susceptibility, Damped Sinusoidal Transients, Cable and Power Leads, 10 kHz to 100 MHz
	RE101	Radiated Emissions, Magnetic Field, 30 Hz to 100 kHz
	RE102	Radiated Emissions, Electric Field, 10 kHz to 18 GHz
	RE103	Radiated Emissions, Antenna Spurious and Harmonic Outputs, 10 kHz to 40 GHz
	RS101	Radiated Susceptibility, Magnetic Field, 30 Hz to 100 kHz
	RS103	Radiated Susceptibility, Electric Field, 2 MHz to 40 GHz RS105 Radiated Susceptibility, Transient Electromagnetic Field

See also https://en.wikipedia.org/wiki/List_of_common_EMC_test_standards

CONTEXT
- Standards, Regulations

- **EMI/EMC Standards may in general be grouped by**
  - EM emissions vs susceptibility to interference (Tx vs Rx)
  - EM coupling mode and/or frequency (conducted, inducted, radiated)
  - Measurement method (depending on i and ii)

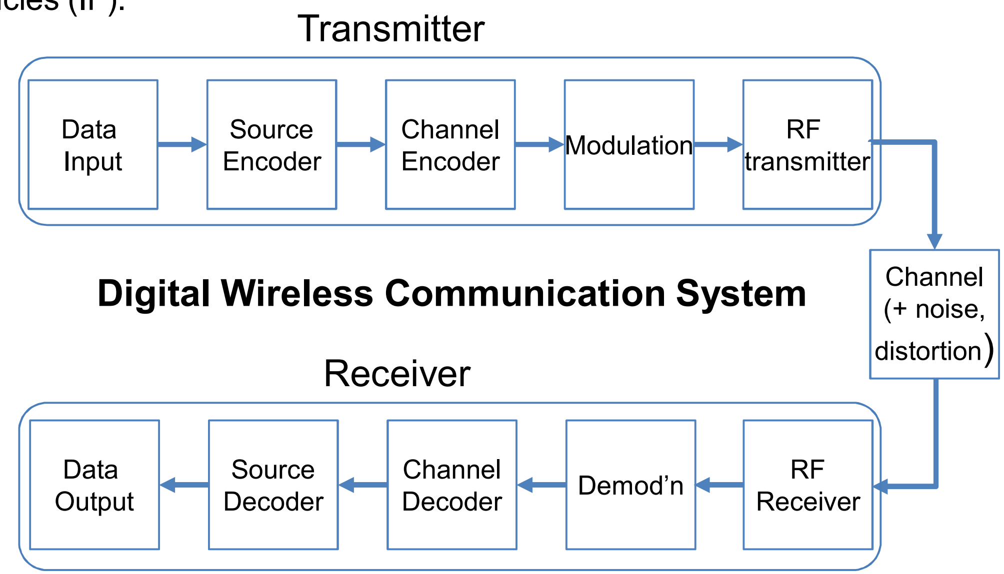
or by

- Equipment function (e.g. power supply, communication)
- Equipment application (e.g. commercial, medical, military)

BACKGROUND
- Digital Communication Systems

The main functional blocks in a digital wireless communication system are as follows…

- **Source Encoding/Decoding**
  Compresses the input data to remove any redundant or unneeded information. For analogue source signals, source coding performs an analogue to digital conversion. Decoding reverses the process.
- **Channel Encoding/Decoding**
  Channel encoding adds some redundancy to minimize the effects of fading and noise in the channel.
- **Modulation/Demodulation**
  The bit stream is modulated to generate the transmitted signal, e.g. pulse modulation at baseband, amplitude and/or phase modulation of RF carrier.
- **Transmission/ Reception**
  Amplification and transduction to/from propagating radio waves. May also involve i) multiplexing for multiple access, ii) mixing to intermediate frequencies (IF).



BACKGROUND
- Potential Impacts of EMI

Susceptibility to interference depends on system design…

- coding
- modulation
- RF carrier frequency, etc.

linked to system requirements…

- data rate
- range (sensitivity)
- multiple access methods
- EM environment
- mobility, etc.

and system application…

- simple wireless point-to-point links (e.g. remote control)
- low power communications (sensing , IoT)
- cellular mobile phone networks (4G → 5G), etc.

BACKGROUND
- Radio Frequency Spectrum

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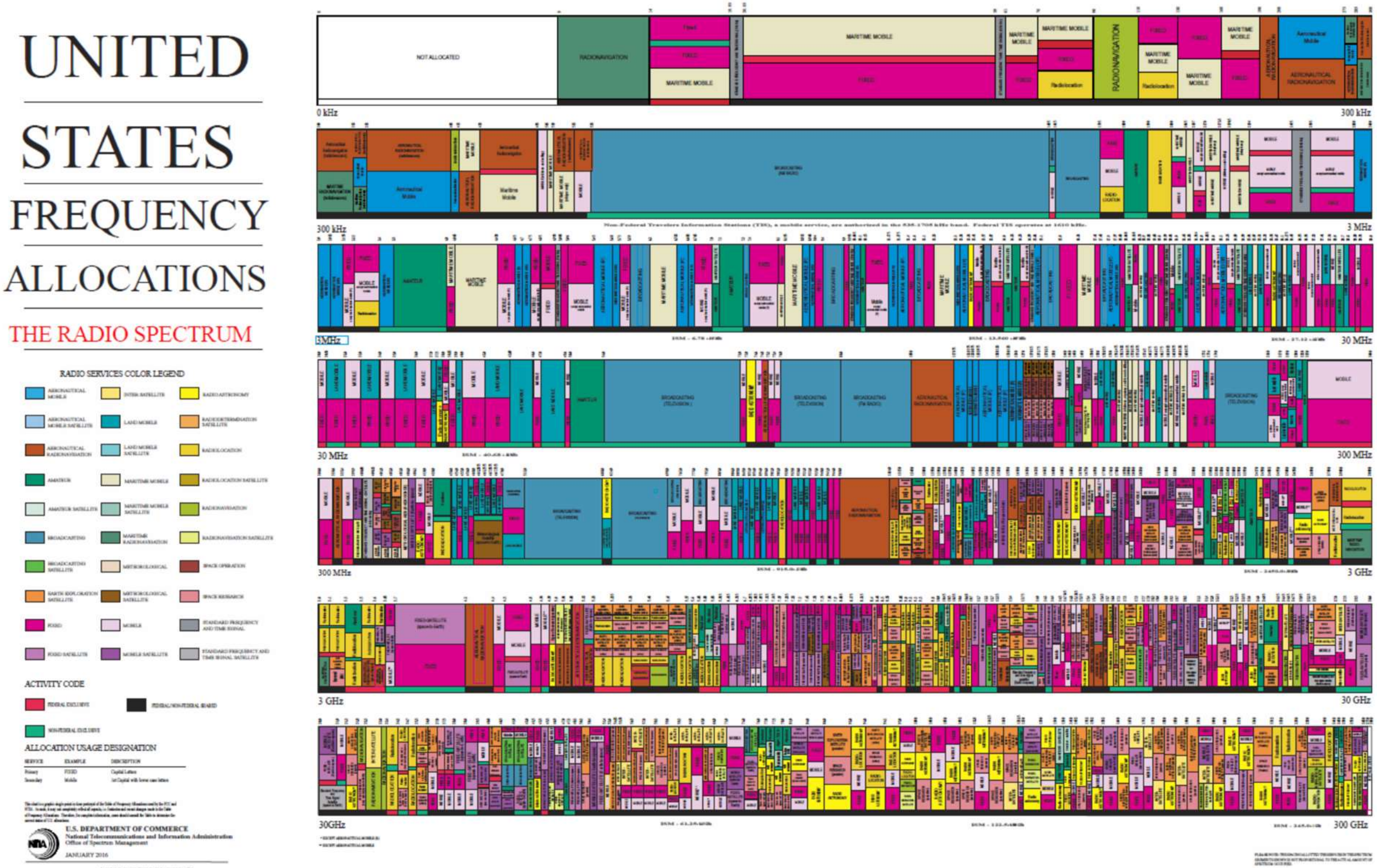


https://www.ntia.doc.gov/files/ntia/publications/january_2016_spectrum_wall_chart.pdf

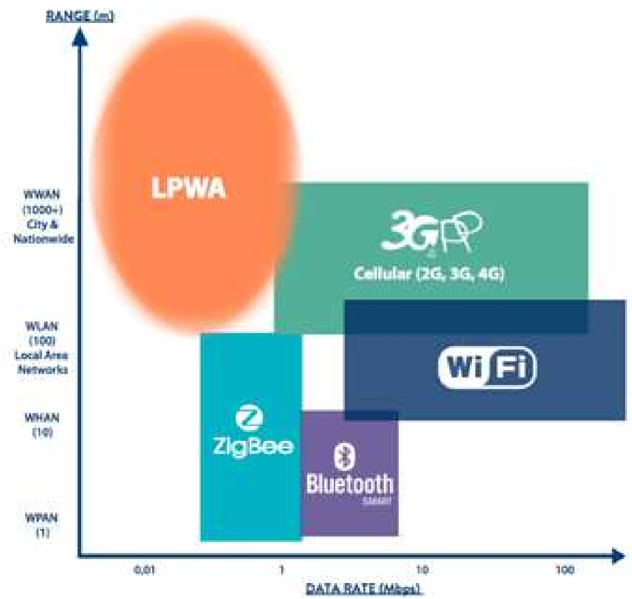
BACKGROUND
- Wireless communication standards

- **Many wireless systems and standards for various communications applications (voice, IoT, etc.)….**

- **Can be classified by…**
  - Range (m to many km)
  - Data rate
  - System complexity
  - License arrangements for use of RF spectrum

- **Examples of wireless communication systems…**
  - WiFi (802.11n)
  - 2G/3G/4G/5G (cellular mobile phone networks, 0.9 – 2.1, 5 GHz)
  - ISM (Industrial, Scientific, Medical) – unlicensed 900MHz, 2.4GHz, etc.
    - Bluetooth, Zigbee (2.4GHz ISM)
  - LPWAN (low power wide area networking), IoT
    - LTE NB (cellular, licensed), LoRa, Sigfox (ISM)

BACKGROUND
- Low-power wide-area network standards



https://www.theiet.org/sectors/information-communications/topics/ubiquitous-computing/articles/lpwan.cfm

	LORA-WAN	SigFox	NB-IoT	CATM1	RPMA
Modulation	DSS with chirp	UNB/ GFSK/ BPSK	OFDMA/ SC-FDMA	OFDMA/ SC-FDMA	RPMA
Frequency	868/ 902-928MHz	868/915 MHz	In band LTE, guard band and stand alone	In band LTE	2.4GHz
Coverage	153-161 dB	149-161 dB	164dB	155.7dB	168-172dB
Bandwidth	125kHz	100Hz (EU)	180kHz	1.08MHz	1MHz
Data rate	0.3 to 50 kbps	100bps	50kbps	1Mbps	624 Kbps DL 156 Kbps UL
Max msg / day	unlimited	140 UL 4 DL	n/a	unlimited	n/a

BACKGROUND
- Narrow-band wireless for IoT

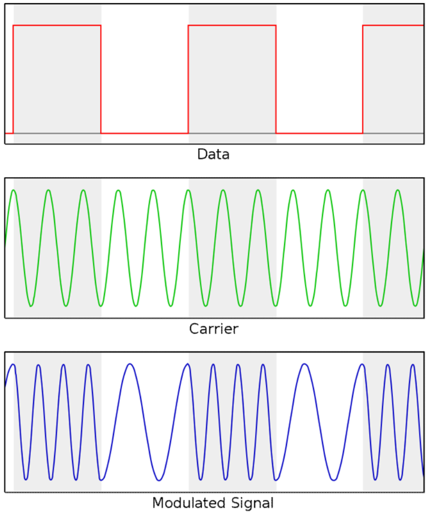
	LTE Cat 1	LTE Cat 0	LTE Cat M1 (eMTC)	LTE Cat NB1 (NB-IoT)
3GPP Release	Release 8	Release 12	Release 13	Release 13
Downlink Peak Rate	10 Mbit/s	1 Mbit/s	1 Mbit/s	250 kbit/s
Uplink Peak Rate	5 Mbit/s	1 Mbit/s	1 Mbit/s	250 kbit/s (multi-tone) 20 kbit/s (single-tone)
Latency	50-100ms	not deployed	10ms-15ms	1.6s-10s
Number of Antennas	2	1	1	1
Duplex Mode	Full Duplex	Full or Half Duplex	Full or Half Duplex	Half Duplex
Device Receive Bandwidth	1.08 - 18 MHz	1.08 - 18 MHz	1.08 MHz	180 kHz
Receiver Chains	2 (MIMO)	1 (SISO)	1 (SISO)	1 (SISO)
Device Transmit Power	23 dBm	23 dBm	20 / 23 dBm	20 / 23 dBm

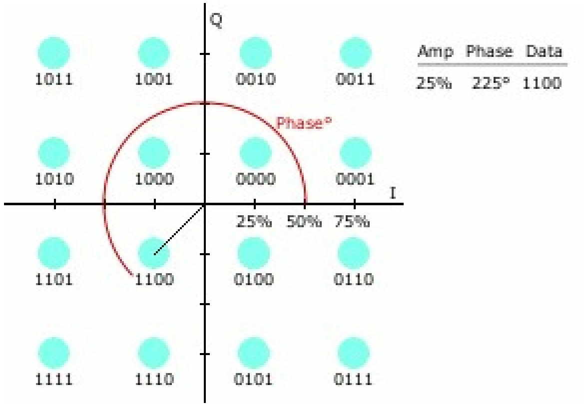
https://en.wikipedia.org/wiki/NarrowBand_IOT

BACKGROUND
- Wireless comunication

RF carrier modulation

- **Analogue (continuous) modulation**
  - Amplitude modulation
  - Frequency/phase modulation
  - Hybrids
- **Digital (quantized, discrete) modulation**
  - ASK
  - FSK
  - PSK
  - Hybrids (e.g. QAM)
- **Spread spectrum methods**
  - Time/Frequency hopping/diversity
  - Direct sequence spread spectrum
  - OFDM





Demodulator designed to distinguish between points in signal constellaion

BACKGROUND
- Potential Impacts of EMI

- **EMI usually regarded as noise in wireless comms.**
- **Effects of noise (interference) depend upon**

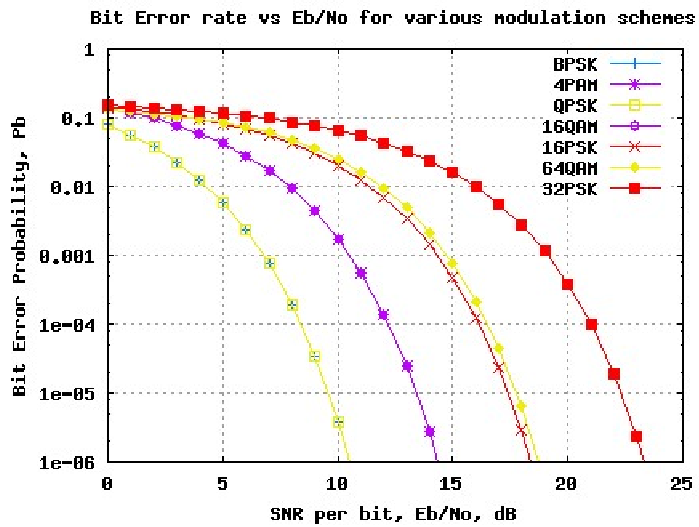
1. Modulation method

- Analogue modulation
  - AM: severe interference, noise = “signal” (e.g. AM radio near computer)
  - F/PM: through nonlinear effects, e.g. overloading of receiver
- Digital modulation
  - Effects of interference generally less obvious, up to a threshold

2. Noise (interference) power relative to signal power

NOTE: “Knee” in BER characteristic…

→ BER deteriorates rapidly with SNR



https://www.embedded.com/print/4017668

BACKGROUND
- Examples of EMI and impact

- Interference between automotobile ignition system and car radio.

  Taggart, Methods of suppressing automotive interference, NBS Special Publication 480-44, 1981. http://nvlpubs.nist.gov/nistpubs/Legacy/SP/nbsspecialpublication480-44.pdf

- Satellite deployment failure due to inductive coupling (crosstalk) between an unshielded attitude control sensor cable and the power bus of the spacecraft. (The control systems cable was redesigned and shielding added).

  Leach et al (NASA), Electronic systems failures and anomalies attributed to electromagnetic interference, NASA-RP-1374, 1995. https://ntrs.nasa.gov/archive/nasa/casi.ntrs.nasa.gov/19960009442.pdf

- Misoperation of neighbour’s garage door opener when LED lighting with dimmer turned on. (Dimmer removed).

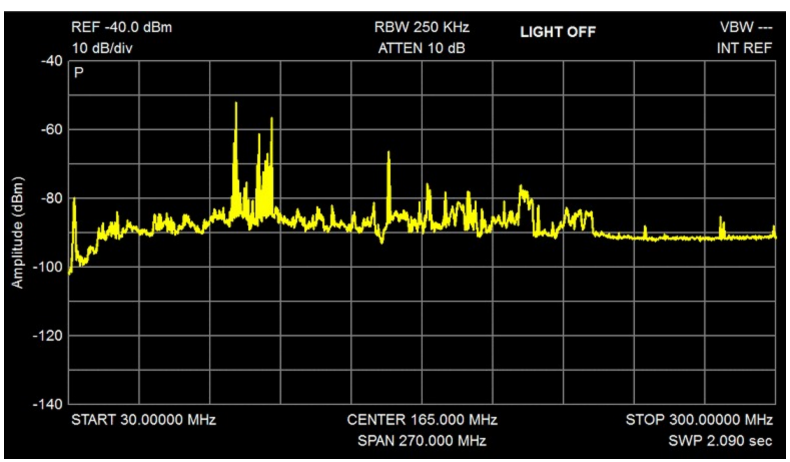
  McHenry, Roberson and Matheson, Electronic Noise is drowning out the Internet of Things, IEEE Spectrum, August 2015. http://spectrum.ieee.org/telecom/wireless/electronic-noise-is-drowning-out-the-internet-of-things

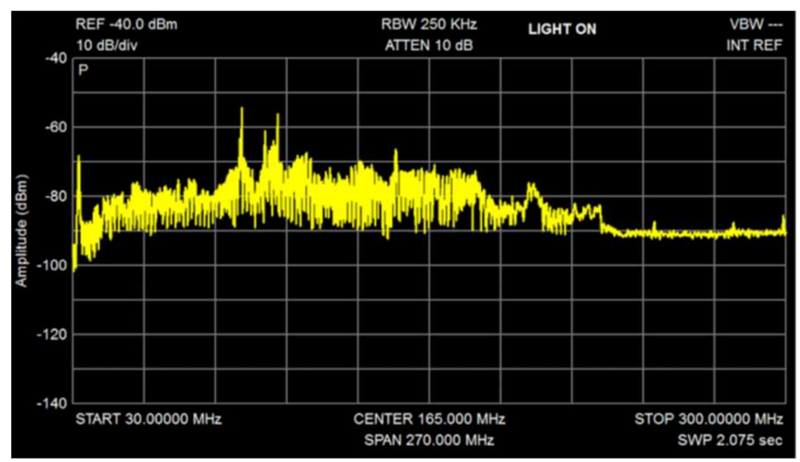
BACKGROUND
- Examples of EMI

5W MR16 LED Downlight Review

General Information

ID	#154
SKU	LTMR5W3K2P
Series	N/A
Manufacturer	Click (other LEDs from Click)
Fitting Type	MR16 (other MR16 LEDs)
Light Type	Downlight (PAR)
Height	48mm
Diameter	PAR16 (50mm)
Weight	42g





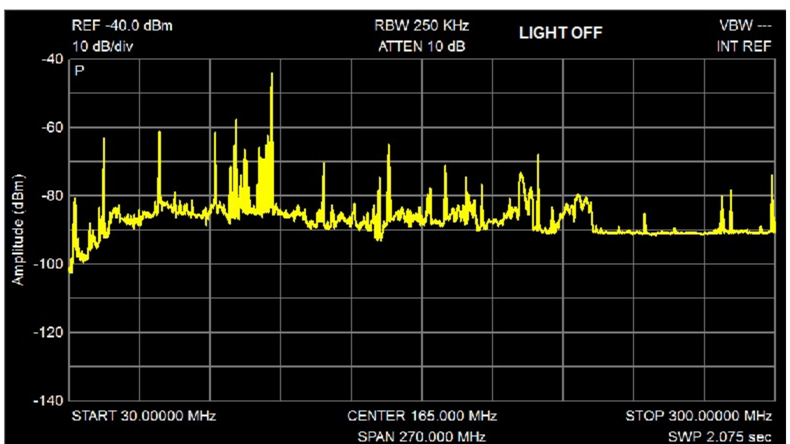
http://www.ledbenchmark.com/faq/LED-interference-issues.html

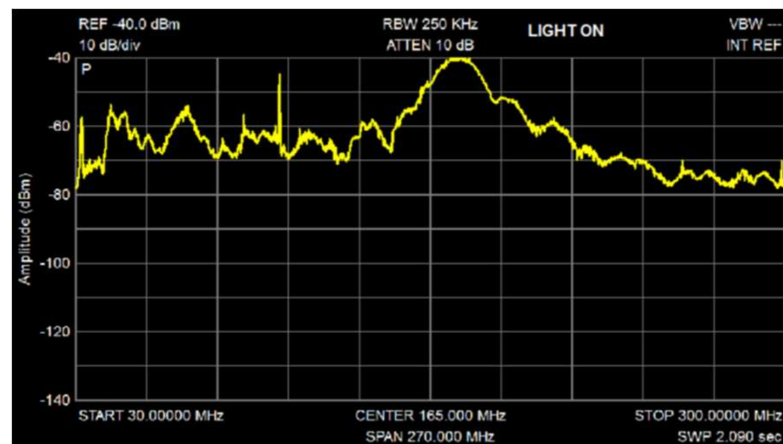
BACKGROUND
- Examples of EMI

5W MR16 Downlight Review

General Information

ID	#131
SKU	LMR16BL27538
Series	N/A
Manufacturer	Mirabella (other LEDs from Mirabella)
Fitting Type	MR16 (other MR16 LEDs)
Light Type	Downlight (PAR)
Height	50mm
Diameter	PAR16 (50mm)
Weight	47g





http://www.ledbenchmark.com/faq/LED-interference-issues.html

PART 2 – Overview
- EMI Fundamentals

- **EMI Fundamentals**
  i) generation
  ii) coupling (energy transfer)

- **Foundations: signals in linear systems**
  i) Signals - two perspectives: time domain ←→ frequency domain
  ii) Systems - two perspectives: electric circuits ←→ electromagnetic fields
  All perspectives are consistent wrt energy in both time and space.

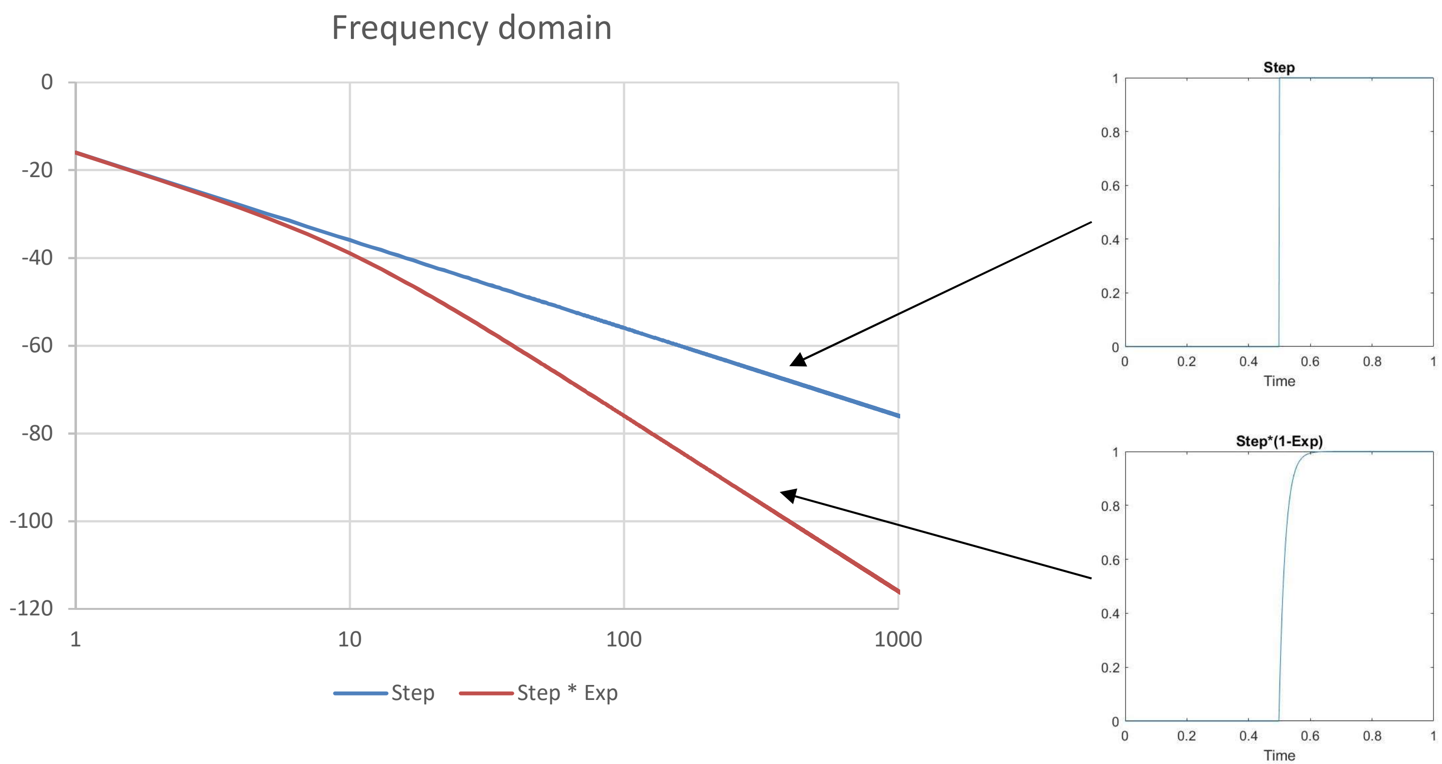
- **Understanding, from an appropriate perspective, how electrical and magnetic energies are <u>embodied</u>, <u>connected</u>, and <u>interact with their environment</u>**
  → effective EMI minimisation and mitigation strategies

Fundamentals of EMI - Signals

- Every switch transition generates di/dt, dv/dt → EMI spectrum

$$\mathcal{F}\{\mathrm{sgn}(t)\} = -j/\omega$$

- ~ ns rise time (GaN) → EMI cutoff frequency ~ 350 MHz



Fundamentals of EMI - Signals

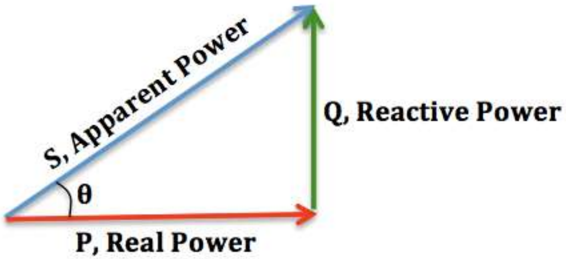
- **Additional material to be included**

    Spectrum of a pulse

    Spectrum of a periodic train of pulses

Fundamentals of EMI
- Electric Circuits

- Differences in electric potential (voltage) drive electrical currents around circuits, transporting electrical energy from source to load.
- Energy *dissipated* in resistive components:
  $R = v_R/i_R$ (resistance, R [Ω] = electric potential [V] per unit current [A])
- Energy *stored* in reactive components:
  $v_L(t) = L.di_L/dt$, $\mathcal{E}_L = ½\, L.i^2$ (inductance L [H] = magnetic flux linkage Λ [Wb] per unit current i [A])
  $i_C(t) = C.dv_C/dt$, $\mathcal{E}_C = ½\, C.v^2$ (capacitance C [F] = electric flux (or charge) ψ [C] per unit voltage v [V])
  Reactance (sinusoidal steady state): $X_L(\omega) = \omega.L$, $X_C(\omega) = -(\omega.C)^{-1}$ (reactance $X(\omega)$ [Ω] = $|V(\omega)/I(\omega)|$)
  Impedance (sinusoidal steady state): $Z = R + jX$ (complex impedance $Z(\omega) = V(\omega)/I(\omega)$ [Ω])
- Power flow (= rate of energy change)
  Instantaneous: $w(t) = v(t).i(t)$
  Sinusoidal steady state: Complex power: $S(\omega) = V(\omega).I^*(\omega) = P + jQ$ [W]
    Real or active power: $P = \mathfrak{Re}\{S\} = S.\cos(\theta)$ [W]
    Reactive power: $Q = \mathfrak{Im}\{S\} = S.\sin(\theta)$ [VAR]
    Apparent power: $S = |S| = (P^2 + Q^2)^{½}$ [VA]
    Power factor: $f = P/S = \cos(\theta)$



- Maximum real power flow from source to load when
  Sinusoidal steady state: $Z_{LD}(\omega) = R_{LD} + j\, X_{LD}(\omega) = R_S - j\, X_S(\omega) = Z_S^*(\omega)$ [NB: when $R_{LD} = R_S$, $X_{LD} = - X_S$]
- Also, waves on transmission lines (distributed circuit; t & d, two conductors)

Fundamentals of EMI - Electromagnetic Fields

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- Energy density at a point in space, **r,** associated with
  - electric fields **E** [V/m] and associated electric flux density **D** = ε.**E** [C/m]
    $\mathcal{E}_E(\mathbf{r}) = \frac{1}{2}\,\varepsilon\,E^2$ [J/m^3], (permittivity $\varepsilon = \varepsilon_r.\,\varepsilon_0$, permittivity of free space ε_0 = 8.85 x10^{-12} [F/m])
  - magnetic fields **H** [A/m] and associated magnetic flux density **B** = μ.**H** [Wb/m]
    $\mathcal{E}_H(\mathbf{r}) = \frac{1}{2}\,\mu\,H^2$ [J/m^3], (permeability $\mu = \mu_r.\,\mu_0$, permeability of free space $\mu_0 = 4\pi$ x10^{-7} [H/m])
  - Changes in E can induce H, and vice versa, even in free space.
  - As for circuits, we may define complex impedance $Z_z(\mathbf{r},\omega) = E_x(\mathbf{r},\omega)/H_y(\mathbf{r},\omega)$ [Ω]
    Note: Orthogonal components of **E** and **H** used. For sinusoidal steady state (at frequency ω), the real (in phase) and imaginary (phase-quadrature) parts related to energy dissipation and storage, respectively.
- Power flow (= rate and direction of energy density change at a point in space)
  Poynting theorem: $\mathbf{P}(\mathbf{r},t) = \mathbf{E}(\mathbf{r},t) \times \mathbf{H}(\mathbf{r},t)$ [W/m^2]
  Sinusoidal steady state: $\mathbf{P}_{av}(\mathbf{r},\omega) + j\mathbf{Q}_{av}(\mathbf{r},\omega) = \frac{1}{2}\,\mathbf{E}_{pk}(\mathbf{r}) \times \mathbf{H}^*_{pk}(\mathbf{r})$ [W/m^2]
- Also, wave phenomena in free space (and in waveguides; t & **r**, single conductor)

NOTE: Field and circuit models consistent - linked by considering energy:

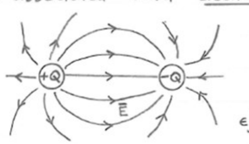
- Voltage difference between two points in space → electric field.
- Moving charge (current) through space → magnetic field.
- Energy stored in static (d/dt = 0) electric and magnetic fields consistent with lumped circuit model
- Electromagnetic wave phenomena in transmission lines consistent with distributed circuit model

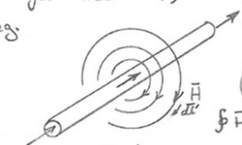
Exception: electromagnetic wave propagation in free space – changing E and H linked by Maxwell, who predicted EM waves before experimental observation – but concept of impedance and other wave phenomena still apply.

Fundamentals of EMI
- Static electromagnetic fields

Below are fundamental concepts in understanding electric + magnetic fields, – they are shown side-by-side to highlight the symmetry in EM field theory.

Electric fields	Magnetic fields
are associated with electric charges	are associated with electric currents (ie. moving electric charges – monopoles not yet discovered)
eg. Q [Coulomb] (↑ symbol, ↖ units); $\epsilon \oiint \bar{E} \cdot d\bar{A} = Q$ (Gauss)	eg. I [Amps] [C s⁻¹]; $\bar{H}$, $d\bar{l}$ (direction of $\bar{H}$ follows R.H.Rule); $\oint \bar{H} \cdot d\bar{l} = I$ (Ampere)
at every point in space, the electric field strength has magnitude and direction ∴ a vector field, denoted **E** or $\bar{E}$, measured in Volts per metre [V m⁻¹]	Magnetic field strength **H** or $\bar{H}$, measured in Amps per metre [A m⁻¹]
Electric charges experience a force in an electric field ($\bar{F} = \bar{E} Q$ [N]) – hence electric currents (moving charge)	Magnetic fields exert a force on moving electric charges ($\bar{F} = \mu dl \bar{I} \times \bar{H}$) – hence electric motors (and generators).
The force can be considered as transmitted by electric flux ψ [C] associated with the electric field by $\bar{D} = \epsilon \bar{E}$ where $\bar{D}$ [C/m²] is electric flux density (ie. p.u. area), and $\epsilon = \epsilon_0 \epsilon_r$ is electric permittivity of the medium ($\epsilon_0 = 8.85 \times 10^{-12}$ [F/m])	The force can be considered as transmitted by magnetic flux ϕ [Wb] associated with the magnetic field by $\bar{B} = \mu \bar{H}$ where $\bar{B}$ [T = Wb/m²] is magnetic flux density (ie. p.u. area), and $\mu = \mu_0 \mu_r$ is magnetic permeability of the medium ($\mu_0 = 4\pi \times 10^{-7}$ [H/m])

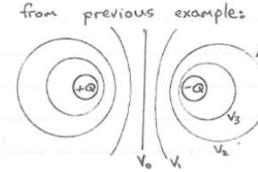




Electric fields	Magnetic fields

Vector fields can be expressed as (i) the gradient of a scalar (potential) field (eg ball rolling down hill – acceleration and direction depends on slope of hill, and height of the hill is proportional to potential energy) plus (ii) the curl of a vector field – Helmholtz theorem.

Electric potential is measured in volts. from previous example:	In regions free of currents + perm. magnets scalar magnetic potential is Φ [A] and $\bar{H} = -\nabla \Phi$ [A m⁻¹]
equipotentials (normal to flux lines) – a scalar field.	In regions not free of currents ($\bar{J} \neq 0$) vector magnetic potential is $\bar{A}$ [Wb m⁻¹] and $\bar{H} = \frac{1}{\mu} \nabla \times \bar{A}$, where $\bar{A} = \int_v \frac{\mu \bar{J}}{4\pi R} dv$
Mathematically $\bar{E} = -\nabla V$ [V m⁻¹] (vector field; gradient operator; scalar field) (eg $\hat{i}\frac{\partial}{\partial x} + \hat{j}\frac{\partial}{\partial y} + \hat{k}\frac{\partial}{\partial z}$)	NB: the definition of magnetic potential is, in general, complicated by the fact that the source of magnetic fields – currents – have direction (ie vector field $\bar{J}$), unlike source of electric fields – charge – (scalar field ρ)



Just as pushing a ball up/down a hill requires/does work (ie. potential energy is gained/lost), moving an electric charge between equipotentials (ie. along force or flux lines) requires/does work. Similarly, moving or rotating current elements in a magnetic field requires/does work.

remember(?) Work [J] = Force [N] × distance [m]

Conversely, the source may be stationary in a changing field.

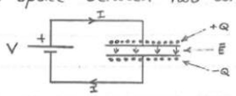
Hence electric and magnetic fields store energy

Energy density (p.u. volume) in an electrostatic field is $dW_E = \frac{1}{2} \epsilon E^2$ [J/m³] ($\epsilon E^2 = \bar{D} \cdot \bar{E}$)	Energy density (p.u. volume) in a magnetostatic field is $dW_M = \frac{1}{2} \mu H^2$ [J/m³]
Total energy stored $W_E = \frac{\epsilon}{2} \int_v E^2 dv$ (volume integral) [J]	Total energy stored $W_m = \frac{\mu}{2} \int_v H^2 dv$ [J]

Fundamentals of EMI
- Electromagnetics of circuits

Lumped circuit elements (static or slowly-varying fields)

A **Capacitor** is a circuit element which stores energy in an electric field in the space between two conductors



defined by $C = \frac{Q}{V}$ [C/V = F] (charge p.u. volt)

eg. If area of plates is S, and separation is d,

$$W_E = \frac{1}{2}\int_v \epsilon E^2 \, dv \qquad \left(E = \frac{V}{d}\right)$$

$$= \frac{1}{2}\epsilon\left(\frac{V}{d}\right)^2 . Sd$$

$$\therefore W_E = \frac{1}{2}\left(\epsilon\frac{S}{d}\right)V^2 = \frac{1}{2}CV^2$$

$$= \frac{1}{2}QV \text{ [J]}$$

where $C \doteq \epsilon\frac{S}{d}$ [F] (ignoring fringe fields)

To change the voltage on the capacitor means a change in electric field, and a change in stored charge (ie current against electric potential = WORK)

now $Q = \int I \, dt = CV$ [C]

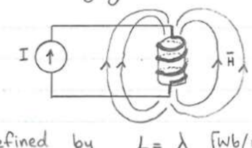
$\therefore$ $\boxed{I = C\frac{dV}{dt}}$ [A]

from cct. theory, work done in charging capacitor from 0 to V

$$W_E = \int VI \, dt \text{ [J]}$$
$$= C\int V dV = \frac{1}{2}CV^2$$

Hence circuit theory and field theory agree on energy stored in a lumped circuit element.

An **Inductor** is a circuit element which stores energy in a magnetic field in the space around a current-carrying wire.

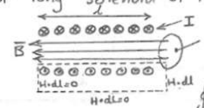


defined by $L = \frac{\lambda}{I}$ [Wb/A = H]

$(\lambda = N\Phi)$ (if no leakage)

$$= \frac{N}{I}\iint_S \bar{B} . d\bar{A} \quad \text{(flux linkage p.u. curr.)}$$

eg. For long solenoid of N turns



Ampere's Law gives..
$$\oint \bar{H} . d\bar{l} = NI$$
$$\therefore H = NI/l \text{ [A/m]}$$

$$W_m = \frac{1}{2}\int_v \mu H^2 \, dv$$

$$= \frac{1}{2}\mu\left(\frac{NI}{l}\right)^2 . Al$$

$$\therefore W_m = \frac{1}{2}\left(\mu\frac{N^2 A}{l}\right)I^2 = \frac{1}{2}LI^2 \text{ [J]}$$

where $L \doteq \mu\frac{N^2 A}{l}$ [H] (ignoring leakage and end-effects)

To change the current flowing in the inductor requires a change in stored magnetic flux

$e = -d\lambda/dt$ (by Faraday)

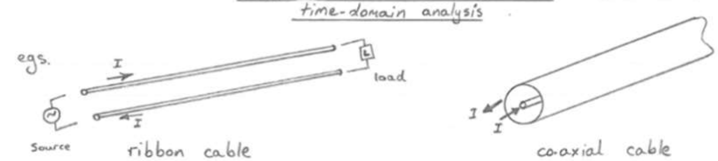
$V = -e$, $\boxed{V = +L\frac{dI}{dt}}$ [V]

From cct. theory, work done in increasing current from 0 to I

$$W_m = \int VI \, dt$$
$$= +L\int I \, dI$$
$$= +\frac{1}{2}LI^2$$

LOSSLESS TRANSMISSION LINES
time-domain analysis

egs.

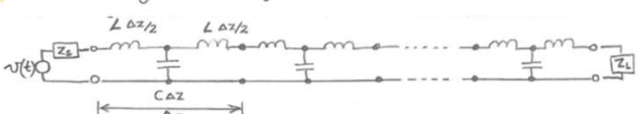


Transmission lines (composed of 2 or more conductors) are useful for transporting electromagnetic energy over a wide range of frequencies (DC – GHz).
[NB: waveguides, composed of single conductors or dielectrics, perform a similar function, but have a minimum usable frequency. Waveguides will be studied in the 2nd half of this course.]

Although the capacitance and inductance of 2-conductor systems can be calculated, for any given length of line, (from electrostatic and magnetostatic theory), how do we model the line?

Except at very low (DC) frequencies, when fields may be considered as "static", neither is correct – we must consider the capacitance and inductance to be distributed along the length of the transmission line.

So



- the smaller Δz, the better the model.
- For linear geometry, we speak of C [F/m], L [H/m].

This approch starts with field theory (statics), but uses circuit theory analysis.

power	50 Hz	6000 km	when wavelength is
telephone	3 kHz	100 km	comparable to size of
FM (TV, radio)	100 MHz	3 m	lumped element (ie line length)
μWave	10 GHz	3 cm	circuit theory breaks down

(roughly, C.T. requires $0{\cdot}1\lambda > z$)

Fundamentals of EMI
- Electromagnetics of distributed circuits

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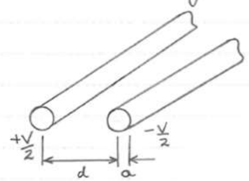
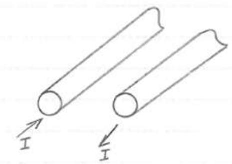
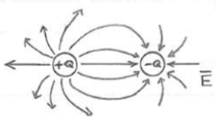
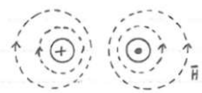
Calculation of typical transmission line parameters - ribbon cable





Simplify geometry - ignore end effects





surface of conductors are equipotentials, and may be considered due to two line-charges.

Calculating potential as function of position (ref Cheng p146-147 RWD p23), for each conductor, and adding, we obtain

$$\boxed{C = \frac{\pi\epsilon}{\cosh^{-1}(d/2a)}} \quad [F/m]$$

If in a medium with finite conductivity,

$$\boxed{G = \frac{\pi\sigma}{\cosh^{-1}(d/2a)}} \quad [S/m]$$

NB conduction current flows along $\bar{E}$, so $\bar{J}$ has same form as $\bar{D}$, hence similar expressions for $C = \frac{Q}{V}$ and $G = \frac{I}{V}$

Approximate analysis:

flux linking one conductor with other

$$\Phi \doteq \int_a^{d-a} \frac{\mu I}{2\pi r} dr = \frac{\mu I}{2\pi} \ln\left(\frac{d-a}{a}\right) \quad [Wb/m]$$

from $\bar{H}$ for single conductor.

Total inductance is

$$L \doteq 2 \cdot \frac{\Phi}{I} = \frac{\mu}{\pi} \ln\left(\frac{d}{a} - 1\right) \quad [H/m]$$

It can be shown exactly (RWD p185) that

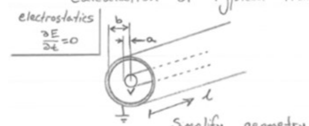
$$\boxed{L = \frac{\mu}{\pi} \cosh^{-1}(d/2a)} \quad [H/m]$$

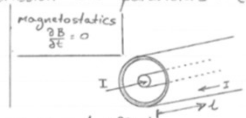
The series resistance depends on skin effect, which causes current to flow near the surface of conductors at high frequencies.

from Cheng p387, RWD p182,

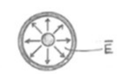
$$\boxed{R = \frac{1}{a}\sqrt{\frac{f\mu}{\sigma\pi}}} \quad [\Omega/m]$$

Calculation of typical transmission line parameters - coaxial cable





Simplify geometry - ignore end effects



Capacitance defined by

$C = Q/V$ [F] where +Q is distributed evenly over surface inner conductor, and -Q is distributed evenly over inner surface of outer conductor.

Taking a cylindrical surface about the inner conductor with radius $a<r<b$

$$\iint_S \bar{D} \cdot d\bar{A} = +Q$$

$$\epsilon E_r \cdot 2\pi r \cdot l = Q$$

$$E_r = \frac{Q}{2\pi\epsilon r l} \quad \text{for } a<r<b \quad [Vm^{-1}]$$

The voltage between inner and outer conductor surfaces is

$$V = \int_a^b E_r \, dr = \frac{Q}{2\pi\epsilon}\left[\ln(r)\right]_a^b = \frac{Q}{2\pi\epsilon} \ln(b/a) \quad [V]$$

substitution gives

$$C = Q/V = \frac{Q}{\frac{Q}{2\pi\epsilon l}\ln(b/a)} = \frac{2\pi\epsilon l}{\ln(b/a)} \quad [F]$$

ie $$\boxed{C = \frac{2\pi\epsilon}{\ln(b/a)} \quad [F/m]}$$

Taking a line integral over a circular path between the conductors

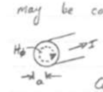
$$I = \oint \bar{H} d\bar{l} = H_\phi 2\pi r$$

$$B_\phi = \mu H_\phi = \frac{\mu I}{2\pi r}$$

$$\Phi_\phi = \int_a^b B \, dr \quad [Wb/m] = \frac{\mu I}{2\pi} \ln(b/a)$$

$$\boxed{L = \frac{\Phi}{I} = \frac{\mu}{2\pi} \ln(b/a)} \quad [H/m]$$

At audio frequencies and lower, the flux inside the centre conductor may be considered



$$J = \frac{I}{\pi a^2} \quad [A/m^2]$$

Current enclosed by path with radius $r<a$, $I(r) = J.\pi r^2$ [A] $= I(r/a)^2$

$$\therefore I(r) = \oint \bar{H}.d\bar{l} = H_\phi . 2\pi r$$

$$\therefore H_\phi = \frac{I.(r/a)^2}{2\pi r} = \frac{I}{2\pi a^2}.r$$

Flux in ring $r+dr$ linking with $I(r)$

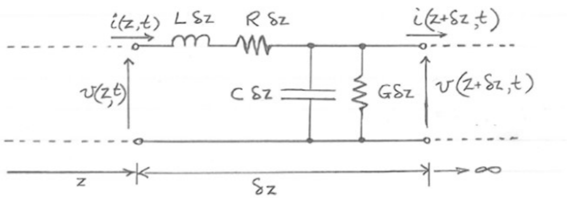
$$d\phi = \mu_0 H_\phi dr = \frac{\mu_w I}{2\pi a^2}.r\,dr$$

$$L_w = \int_0^I \frac{I(r)}{I} d\phi = \int_0^a \left(\frac{r}{a}\right)^2 \cdot \frac{\mu r}{2\pi a^2} dr$$

$$\therefore L_w = \frac{\mu_w}{8\pi} \; [H/m], \quad L_{tot} = \frac{\mu_w}{8\pi} + \frac{\mu}{2\pi}\ln(b/a)$$

eg. RG58 coax $a = 0.45$ mm, $b = 1.5$ mm, $\epsilon_r = 2.28$, $\mu_r = 1$ $\therefore \epsilon = 1.95\times10^{-11}$ [F/m] $\therefore C = 100$ [pF/m], $L = 240$ [nH/m]

Fundamentals of EMI
- Waves on transmission lines



[L, R, C, G in [units/m]
Lδz, Rδz, Cδz, Gδz, look like lumped elements]

by KVL: $\delta v(z,t) = v(z+\delta z, t) - v(z,t) = -L\delta z . \frac{\partial i(z,t)}{\partial t} - R\delta z . i(z,t)$

$$\frac{\delta v}{\delta z} = -L\frac{\partial i}{\partial t} - Ri \qquad (1)$$

by KCL: $\delta i(z,t) = i(z+\delta z,t) - i(z,t)$

$$= -v(z+\delta z,t).G\,\delta z - C\delta z . \frac{\partial v(z+\delta z,t)}{\partial t}$$

$$\frac{\delta i}{\delta z} = -C\frac{\partial v}{\partial t} - Gv \qquad (2)$$

In the limit $\delta z \to 0$, (1) and (2) become

$$-\frac{\partial v}{\partial z} = L\frac{\partial i}{\partial t} + Ri$$

$$-\frac{\partial i}{\partial z} = C\frac{\partial v}{\partial t} + Gv$$

In many cases, the losses due to R and G can be neglected. This simplifies analysis for non-sinusoidal signals, so

$$\boxed{\frac{\partial v}{\partial z} = -L\frac{\partial i}{\partial t} \quad (3) \qquad \frac{\partial i}{\partial z} = -C\frac{\partial v}{\partial t} \quad (4)}$$

"Telegraphers' Eqns" (derived by Heaviside about 1860.)

$\frac{\partial}{\partial t}(3) \to \frac{\partial^2 v}{\partial t\,\partial z} = -L\frac{\partial^2 i}{\partial t^2}$

$\frac{\partial}{\partial z}(4) \to \frac{\partial^2 i}{\partial z^2} = -C\frac{\partial^2 v}{\partial z\,\partial t}$

$$\frac{\partial^2 i}{\partial z^2} = LC\frac{\partial^2 i}{\partial t^2} \qquad (5)$$

and similarly, $\frac{\partial^2 v}{\partial z^2} = LC\frac{\partial^2 v}{\partial t^2} \qquad (6)$

These are wave equations,

so called, because the solution to this form of DE are wave functions:

$$v(z,t) = v_f\left(t-\frac{z}{v_p}\right) + v_b\left(t+\frac{z}{v_p}\right) \qquad (7)$$

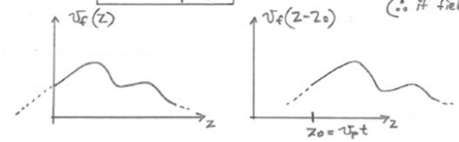
$$i(z,t) = i_f\left(t-\frac{z}{v_p}\right) + i_b\left(t+\frac{z}{v_p}\right) \qquad (8)$$

$$= \frac{1}{Z_0}\left\{v_f\left(t-\frac{z}{v_p}\right) - v_b\left(t-\frac{z}{v_p}\right)\right\}, \quad Z_0 = \sqrt{L/C}$$

where subscript "f" is the forward (→ +ve z) travelling wave,
" "b" " " backward (−ve z ←) " " ,

and $\boxed{v_p = 1/\sqrt{LC}}$ is the phase velocity of the waves [m/s].
(∵ if fields between conductors only, $LC = \mu\varepsilon$)

eg.



[the wave shape (Voltage distribution) is maintained, but shifts in z with t.]

(eg RG58 coax $v_p = 2\times10^8$ [m/s], $1/v_p = 5$ [ns/m])

Now let's consider the ratio v/i:

from (3) $\frac{\partial v}{\partial z} = -\frac{1}{v_p}v_f' + \frac{1}{v_p}v_b' = -L(i_f' + i_b')$

from (4) $\frac{\partial i}{\partial z} = -\frac{1}{v_p}i_f' + \frac{1}{v_p}i_b' = -C(v_f' + v_b')$

$(10)\times -\frac{1}{C} + (9)\times -v_p \Rightarrow 2v_f' = \left(\frac{1}{Cv_p} + Lv_p\right)i_f' + \left(Lv_p - \frac{1}{Cv_p}\right)^{0} i_b'$

$$= 2\sqrt{\frac{L}{C}}\, i_f'$$

$\therefore \frac{dv_f}{di_f} = \sqrt{\frac{L}{C}}$ [Ω] (also $\frac{dv_b}{di_b} = -\sqrt{\frac{L}{C}}$)

$\sqrt{L/C}$ has dimension of impedance, and is defined as the characteristic impedance of the transmission line

$$\boxed{Z_0 = \sqrt{L/C} \ [\Omega]}$$

(eg RG58 coax $Z_0 = \sqrt{\frac{240\times10^{-9}}{100\times10^{-12}}} = 49$ [Ω])

Fundamentals of EM waves
- Reflections on transmission lines

The question now arises – what happens on a finite length T.L. when the forward travelling wave comes to the end of the line (ie. the load)?

Basically, this is a boundary value problem – the load impedance Z_L constrains the v/i ratio at the end of the line.

1) If $Z_L = Z_0$, then the load looks like an infinitely long transmission line (of the same characteristic impedance) and no reflections occur.

2) If $Z_L \neq Z_0$, then $v_L/i_L \neq v_f/i_f$ and a reflected wave is set up so that the total voltage and current waves satisfy $\boxed{\dfrac{v_f + v_b}{i_f + i_b} = Z_L}$.

The forward and backward waves are still defined by $\dfrac{v_f}{i_f} = -\dfrac{v_b}{i_b} = Z_0$

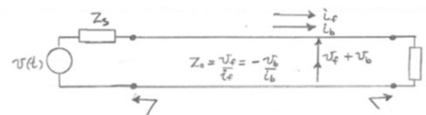
Therefore, $\dfrac{v_f + v_b}{v_f - v_b} = \dfrac{Z_L}{Z_0}$

Alternatively, the magnitude of the reflected wave relative to the incident wave is:

$$\boxed{\rho = \frac{v_b}{v_f} = -\frac{i_b}{i_f} = \frac{Z_L - Z_0}{Z_L + Z_0}}$$

ρ is defined as the reflection coefficient

[NB ρ may be complex, implying magnitude and phase differences in the reflected wave, if Z_L complex. If $Z_L = Z_0$, $\rho = 0$, ie no reflection occurs.]



$\rho_s = \dfrac{Z_s - Z_0}{Z_s + Z_0}$

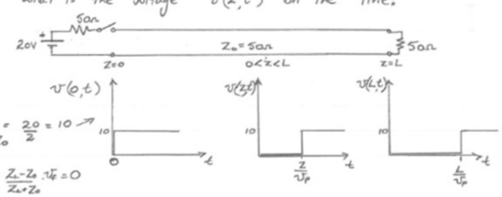
$\dfrac{v_f(l,t) + v_b(l,t)}{i_f(l,t) + i_b(l,t)} = \dfrac{v_L(t)}{i_L(t)} = Z_L$

$\dfrac{v_b(l,t)}{v_f(l,t)} = \rho_L = \dfrac{Z_L - Z_0}{Z_L + Z_0}$

Eg 1 A switch connects a 20 V_{DC} source to a T.L. with $Z_s = Z_L = Z_0 = 50\Omega$, $v_p = 2\times10^8$ m/s at time $t = 0$. What is the voltage $v(z,t)$ on the line?

$v_f = V_s \dfrac{Z_0}{Z_s + Z_0} = \dfrac{20}{2} = 10$

$v_b = \rho_L v_f = \dfrac{Z_L - Z_0}{Z_L + Z_0} v_f = 0$



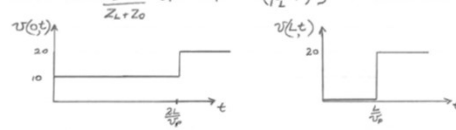
Eg 2 As above, except $Z_L = \infty$ (ie. open circuit at load)

$v_f = V_s \dfrac{Z_0}{Z_s + Z_0} = 10V$ as above

$v_b = \dfrac{Z_L - Z_0}{Z_L + Z_0} v_f = v_f$ $(\rho_L = 1)$

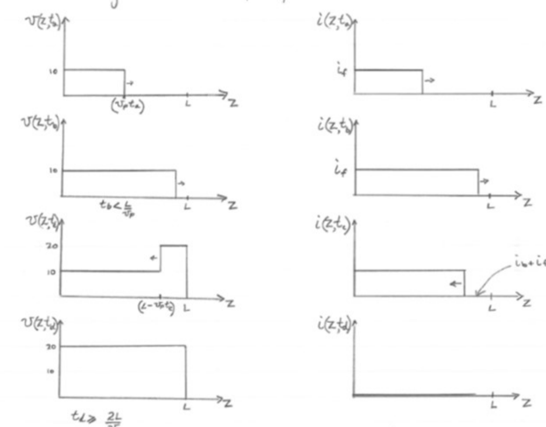
$v_L = v_b + v_f = 2v_f$

$i_L = i_b + i_f = 0$



NB if the source impedance was not "matched" to the transmission line (ie if $Z_s \neq Z_0$), then the backward wave v_b would be reflected from source, giving another forward wave, $v_{f_2} = \left(\dfrac{Z_s - Z_0}{Z_s + Z_0} + 1\right) v_{b_1} + v_{f_1}$

Could draw $v(z)$, $i(z)$ at various instants in time to get another perspective:



$i_f = \dfrac{V_s}{Z_0 + Z_s} = \dfrac{20}{100} = 0.2$ A

$i_b = -i_f$

Fundamentals of EM waves
- Sinusoidal waves on transmission lines

We will now derive the above using phasors, and include the effect of losses.

now
$$v(z,t) = \mathrm{Re}\{V(z)\, e^{j\omega t}\} \quad (1)$$
$$i(z,t) = \mathrm{Re}\{I(z)\, e^{j\omega t}\} \quad (2)$$

and the Telegrapher's Eqns, with loss, are

$$-\frac{\partial v}{\partial z} = Ri + L\frac{\partial i}{\partial t} \quad (3)$$
$$-\frac{\partial i}{\partial z} = Gv + C\frac{\partial v}{\partial t} \quad (4)$$

$$\frac{\partial^2 V(z)}{\partial z^2} = +(R+j\omega L)(G+j\omega C).V(z) = \gamma^2 V(z) \quad (7)$$
$$\frac{\partial^2 I(z)}{\partial z^2} = +(G+j\omega C)(R+j\omega L).I(z) = \gamma^2 I(z) \quad (8)$$

(7) and (8) may easily be solved, to give

$$V(z) = V_f(z) + V_b(z) = V_{fo}\, e^{-\gamma z} + V_{bo}\, e^{\gamma z} \quad (9)$$
$$I(z) = I_f(z) + I_b(z) = I_{fo}\, e^{-\gamma z} + I_{bo}\, e^{\gamma z} \quad (10)$$

Here γ is called the propagation constant where

$$\gamma = \alpha + j\beta = \sqrt{(R+j\omega L)(G+j\omega C)} \quad (11)$$

α: attenuation constant [Np/m]; β: phase constant [rad/m]

$\beta = \omega/v_p$ (how fast phase changes) $\left[\frac{r/s}{m/s}\right] = \left[\frac{r}{m}\right]$ (spatial frequency)

The characteristic impedance is defined, as before, by

$$Z_0 = \frac{V_f}{I_f} = -\frac{V_b}{I_b} = \sqrt{\frac{R+j\omega L}{G+j\omega C}} \ [\Omega] \quad \left(\text{for a lossless line, } Z_0 = \sqrt{L/C}\right)$$

where R, G, L, C, are again in [units/metre] (12)

Example

A telephone line is measured at 1kz, and found to have $R = 4.2\ \Omega/\mathrm{km}$, $G = 0.2\ \mu S/\mathrm{km}$, $L = 2.2\ \mathrm{mH/km}$, $C = 5.4\ \mathrm{nF/km}$

Given $v(0,t) = \cos(\omega t)$, if frequency is 1kHz, calculate characteristic impedance, attenuation, and delay at a point 10km along the line. Also write both the time domain and phasor formulae for voltage + current there.

Soln: $\omega = 2\pi f = 2\pi \times 10^3$ [r/s]

from (12), $$Z_0(1\,\mathrm{kHz}) = \sqrt{\frac{4.2 + j\times 2\pi\times 10^3 \times 2.2\times 10^{-3}}{0.2\times 10^{-6} + j\times 2\pi \times 10^3 \times 5.4\times 10^{-9}}} = \sqrt{\frac{4.2 + j.13.8}{0.2\times 10^{-6} + j\,3.4\times 10^{-5}}}$$

$$\therefore Z_0(1\,\mathrm{kHz}) = 646 - j94\ [\Omega] = 652.8 \angle -8.3^\circ$$

from (11) $$\gamma(1\,\mathrm{kHz}) = \sqrt{(4.2 + j13.8)(0.2\times 10^{-6} + j\,3.4\times 10^{-5})} = \sqrt{(-0.47 + j0.15)\times 10^{-3}}$$

$$\therefore \gamma(1\,\mathrm{kHz}) = 0.0033 + j0.022\ [\mathrm{km}^{-1}]$$

$$\therefore \alpha(1\,\mathrm{kHz}) = 0.0033\ [\mathrm{Np/km}]$$
$$\beta(1\,\mathrm{kHz}) = 0.022\ [\mathrm{rad/km}]$$

10km down line:

$$\alpha = 0.033\ [\mathrm{Np}], \ \therefore V_f = V_{fo}.e^{-0.033} = 0.967\, V_{fo}\ [\mathrm{Volt}]$$
$$\tau_d = L/v_p = L\frac{\beta}{\omega} = 10 \times \frac{0.022}{2\pi\times 10^3} = 35\ [\mu s]$$

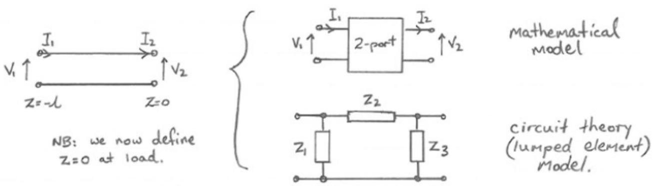
$$v_f(10\mathrm{km}, t) = 0.967\cos(2\pi\times 10^3 t - 0.22^c), \quad V_f(10\mathrm{km}) = 0.967\, e^{-j0.22^c}$$

$$I_f = \frac{V_f}{Z_0} = \frac{0.967\, e^{-j0.22}}{644 - j93} = 1.5\angle -4.4^\circ\ [\mathrm{mA}], \quad i_f = 1.5\cos(2\pi\times 10^3 t - 0.076^c)\ [\mathrm{mA}]$$

Fundamentals of EM waves
- Impedance transformations

We are often only interested in what happens at the terminals (ie input + output) of a transmission line, rather than what goes on in between (which is usually inaccessible anyway).

In such situations, it is useful to have a model for the transmission line



Using these models we will see how transmission lines can be used for <u>impedance transformation</u>, and how they can be used as <u>lumped elements</u>, at any <u>one frequency</u>.

<u>2-Port model</u>

In general, any circuit with an input and output can be modelled as a 2-port network ie described by parameters (usually frequency dependent) which relate the input + output voltages + currents.

For transmission lines, transmission parameters are best

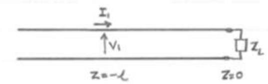
$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix} \quad (14)$$

i/p — xmission matrix — o/p

Moving back towards the source, the total voltage and current at $z=-l$ is

$$(17)\quad V_1 = V_f e^{\gamma l} + V_b e^{-\gamma l}$$

$$(18)\quad I_1 = \frac{V_f}{Z_0} e^{\gamma l} - \frac{V_b}{Z_0} e^{-\gamma l}$$



Solving (15) and (16) for V_f and V_b, and subst. into (17) and (18),

$$V_f = \frac{1}{2}(V_2 + Z_0 I_2)$$

$$V_b = \frac{1}{2}(V_2 - Z_0 I_2)$$

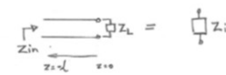
$$\therefore\ V_1 = V_2 \cdot \frac{e^{\gamma l} + e^{-\gamma l}}{2} + Z_0 I_2 \frac{e^{\gamma l} - e^{-\gamma l}}{2} = V_2 \cosh \gamma l + Z_0 I_2 \sinh \gamma l$$

$$I_1 = \frac{V_2}{Z_0} \frac{e^{\gamma l} - e^{-\gamma l}}{2} + I_2 \cdot \frac{e^{\gamma l} + e^{-\gamma l}}{2} = \frac{V_2 \sinh \gamma l}{Z_0} + I_2 \cosh \gamma l$$

<u>ie</u>

$$\boxed{\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \cosh \gamma l & Z_0 \sinh \gamma l \\ \frac{1}{Z_0} \sinh \gamma l & \cosh \gamma l \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}} \quad (19)$$

Using the mathematical model we can calculate the impedance seen looking in to the transmission line at any point, given a load Z_L



Impedance seen at $z=-l$ is:

from (19)

$$(20)\quad Z_{in} = \frac{V_1}{I_1} = \frac{A V_2 + B V_2/Z_L}{C V_2 + D V_2/Z_L} = Z_0 \left(\frac{Z_L \cosh \gamma l + Z_0 \sinh \gamma l}{Z_L \sinh \gamma l + Z_0 \cosh \gamma l} \right)$$

<u>Alteratively</u> equations (17) + (18) could be written using $\rho = \frac{V_b}{V_f} = (Z_L - Z_0)/(Z_L + Z_0)$

$$V_1 = V_f e^{+\gamma l} (1 + \rho e^{-2\gamma l})$$

$$I_1 = \frac{V_f}{Z_0} e^{\gamma l} (1 - \rho e^{-2\gamma l})$$

so $\boxed{Z_{in} = Z_0 \frac{1 + \rho e^{-2\gamma l}}{1 - \rho e^{-2\gamma l}}}$ (21)

<u>The load impedance Z_L is transformed to look like Z_{in} by T.L.</u>

Fundamentals of EMI
- Electromagnetic fields

- **Maxwells equations (point or differential form)**
  - Describe how EM fields vary about a point in space, **r**

At any point, E-M fields obey the following laws.

① $\nabla \cdot \bar{D}(\bar{r},t) = \rho(\bar{r},t)$ Gauss' Law

② $\nabla \cdot \bar{B}(\bar{r},t) = 0$

③ $\nabla \times \bar{E}(\bar{r},t) = -\frac{\partial \bar{B}(\bar{r},t)}{\partial t}$ Faraday Law

④ $\nabla \times \bar{H}(\bar{r},t) = \bar{J}(\bar{r},t) + \frac{\partial \bar{D}(\bar{r},t)}{\partial t}$ Ampere-Maxwell Law

- **Constitutive equations**
  - Describe the medium in which the fields exist

⑤ $\bar{D}(\bar{r},t) = \bar{\bar{\epsilon}}(\bar{r},\bar{E})\,\bar{E}(\bar{r},t)$

⑥ $\bar{B}(\bar{r},t) = \bar{\bar{\mu}}(\bar{r},\bar{H})\,\bar{H}(\bar{r},t)$

⑦ $\bar{J}(\bar{r},t) = \bar{\bar{\sigma}}(\bar{r},\bar{E})\,\bar{E}(\bar{r},t)$

Note: in general, all variables

- are vectors – they have direction and magnitude
- represent fields – distributed throughout space, function of **r**

Fundamentals of EMI
- Electromagnetic fields and waves

All engineering electromagnetic phenomena can be described by solving the partial differential equations 1 – 4, subject to

i) constitutive equations 5-7, and

ii) boundary conditions (depend on physical layout, to be specified).

Solutions of Maxwell's equations predict to a high degree of accuracy phenomena such as

- Electromagnetic coupling in and between "lumped" electrical circuits
- Behaviour of distributed electrical circuits (dimensions comparable to λ)
- Electromagnetic wave propagation and interaction with materials
- Transduction between circuits and EM waves (antennas)
- Guidance of EM waves
- EM waves in free space

Fundamentals of EMI
- Electromagnetic fields and waves

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- **Meaning of symbols, ISI units**

Symbols [units]

$\bar{r}$ [m, radians] A vector which denotes position in space relative to some reference point. The coordinate system used may be any one of the orthogonal coordinate systems (rectangular, cylindrical, spherical, etc) but should be chosen to use symmetries present in the boundary conditions.

$\bar{D}$ [$C.m^{-2}$] Electric flux density, a vector field. ie (at any instant in time) the flux density may have a unique direction and magnitude at every point in space.

ρ [$C.m^{-3}$] Charge density, a scalar field ie (at any instant in time) the charge density may have a unique magnitude at every point in space. (density does not have direction)

$\bar{E}$ [$V.m^{-1}$] Electric field strength, a vector field

$\bar{B}$ [$T = Wb.m^{-2}$] Magnetic flux density, a vector field

$\bar{H}$ [$A\,m^{-1}$] Magnetic field strength, a vector field

$\bar{J}$ [$A.m^{-2}$] Current density, a vector field

∇ gradient operator "del", a vector operator

eg in 3D rectilinear coordinates $\nabla \equiv \bar{i}\frac{\partial}{\partial x} + \bar{j}\frac{\partial}{\partial y} + \bar{k}\frac{\partial}{\partial z}$

where $\bar{i}, \bar{j}, \bar{k}$ are unit vectors in the x, y, z directions, respectively.

The gradient of a scalar function (eg $\nabla\phi$) (at any point) is a vector which points in the direction of greatest change of ϕ (eg if $\phi(\bar{r})$ represents the elevation of a hill as a function of position, then a ball placed at $\bar{r}$ on the hill will roll in the direction of $-\nabla\phi$).

Fundamentals of EMI
- Electromagnetic fields

- **Meaning of symbols, vector operators**

$\nabla \cdot \bar{F}$ divergence, scalar product of gradient and vector field

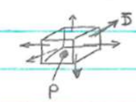
eg in 3D rectilinear coordinates

$$\nabla \cdot \bar{F} = \left(\bar{x}\frac{\partial}{\partial x} + \bar{y}\frac{\partial}{\partial y} + \bar{z}\frac{\partial}{\partial z}\right) \cdot \left(\bar{x}F_x + \bar{y}F_y + \bar{z}F_z\right)$$
$$= \frac{\partial F_x}{\partial x} + \frac{\partial F_y}{\partial y} + \frac{\partial F_z}{\partial z}$$

The divergence of a vector field (at any point) gives a scalar measure of flux flow per unit volume, or equivalently, source strength per unit volume at that point. (eg $\nabla \cdot \bar{D} = \rho$) or field variation in the direction of flow.

The divergence theorem relates the net flux flowing through a closed surface to the source strength in the enclosed finite volume

(9) $$\oint_{S(V)} \bar{F} \cdot d\bar{S} = \int_{V(S)} \nabla \cdot \bar{F} \, dV$$

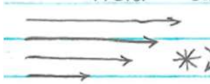


$\nabla \times \bar{F}$ curl, vector product of gradient and vector field.

eg in 3D rectilinear coordinates

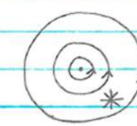
$$\nabla \times \bar{F} = \begin{vmatrix} \bar{x} & \bar{y} & \bar{z} \\ \partial/\partial x & \partial/\partial y & \partial/\partial z \\ F_x & F_y & F_z \end{vmatrix} = \bar{x}\left(\frac{\partial}{\partial y}F_z - \frac{\partial}{\partial z}F_y\right) + \bar{y}\left(\frac{\partial}{\partial z}F_x - \frac{\partial}{\partial x}F_z\right) + \bar{z}\left(\frac{\partial}{\partial x}F_y - \frac{\partial}{\partial y}F_x\right)$$

The curl of a vector field (at any point) gives a vector measure of the line integral per unit area enclosed. It is NOT a measure of field curvature, but a measure of field variation across its direction of flow. The rotational velocity of an infinitesimally small paddle wheel placed at some point in a field measures the curl of the field at that point.



$\nabla \times \bar{F} \neq 0$

eg water flow over a surface

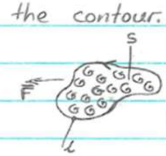


$\nabla \times \bar{H} = 0$

eg magnetic flux around a constant line current

Stokes theorem relates the line integral of field around a closed contour (of finite length) to the total curl of the field over any surface enclosed by the contour.

(10) $$\oint_{l(A)} \bar{F} \cdot d\bar{l} = \int_{S(l)} \nabla \times \bar{F} \cdot d\bar{S}$$



Fundamentals of EMI
- Electromagnetic fields

Special situations leading to simplification of equations

1) Maxwell's equations
   a) static fields
   b) sinusoidal (time-harmonic) fields
   c) charge free, current free regions

1a) if the EM fields depend only on position, that is they are "static" in time, eqns ① to ④ become

$$\nabla\cdot\bar{D} = \rho \qquad \nabla\times\bar{E} = 0$$
$$\nabla\cdot\bar{B} = 0 \qquad \nabla\times\bar{H} = \bar{J}$$

(Students should be familiar with static electric and magnetic fields)

b) if the EM fields vary sinusoidally in time (in a stationary frame)

ie $\bar{E}(\bar{r},t) = Re\{\hat{E}(\bar{r})\, e^{j\omega t}\}$ $\quad \bar{J}(\bar{r},t) = Re\{\hat{J}(\bar{r})\, e^{j\omega t}\}$

$\bar{B}(\bar{r},t) = Re\{\hat{B}(\bar{r})\, e^{j\omega t}\}$ $\quad \rho(\bar{r},t) = Re\{\hat{\rho}(\bar{r})\, e^{j\omega t}\}$

equations ① to ④ may be written

$$\nabla\cdot\hat{D} = \hat{\rho} \qquad \nabla\times\hat{E} = -j\omega\hat{B}$$
$$\nabla\cdot\hat{B} = 0 \qquad \nabla\times\hat{H} = \hat{J} + j\omega\hat{D}$$

these are the complex time harmonic form of Maxwell's eqns.

c) i) in a charge free region $\rho(\bar{r},t) = 0$

ii) in a region free of conduction current $\bar{J}(\bar{r},t) = 0$

In free space, both (i) and (ii) are true so,

$$\nabla\cdot\bar{D} = 0 \qquad \nabla\times\bar{E} = -\partial\bar{B}/\partial t$$
$$\nabla\cdot\bar{B} = 0 \qquad \nabla\times\bar{H} = \partial\bar{D}/\partial t$$

(note the symmetry of Maxwell's eqn's in free space)

2) Constitutive equations
   a) isotropic media
   b) homogeneous media
   c) linear media

2 a) A medium is isotropic if it has the same property in all directions (ie scalar), then eqns ⑤ and/or ⑥ and/or ⑦ become

⑤a $\bar{D}(\bar{r},t) = \epsilon(\bar{r},E)\,\bar{E}(\bar{r},t)$

⑥a $\bar{B}(\bar{r},t) = \mu(\bar{r},H)\,\bar{H}(\bar{r},t)$

⑦a $\bar{J}(\bar{r},t) = \sigma(\bar{r},E)\,\bar{E}(\bar{r},t)$

Anisotropic media are used in devices such as light polarisers and circulators, however we won't be studying such devices in this course, so from now on totally isotropic media will be assumed. (It is rare that more than one of ϵ, μ, σ be anistropic in any case)

b) A medium is homogeneous if it has the same property at all points (ie independent of $\bar{r}$).
Therefore for homogeneous and isotropic media equations ⑤a and/or ⑥a and/or ⑦a become

⑤b $\bar{D}(\bar{r},t) = \epsilon(E)\,\bar{E}(\bar{r},t)$

⑥b $\bar{B}(\bar{r},t) = \mu(H)\,\bar{H}(\bar{r},t)$

⑦b $\bar{J}(\bar{r},t) = \sigma(E)\,\bar{E}(\bar{r},t)$

Ideal media are generally regarded as homogeneous unless the inhomogeneity is purposefully controlled (eg by modifying a homogeneous material with the inclusion of material with different properties)
An example of an inhomogeneous medium is the ionosphere.
All media considered in this course will be homogeneous.

c) A medium is linear if its property is independent of the applied field (ie constant wrt. $E = |\bar{E}|$, $H = |\bar{H}|$)
Equations ⑤b and/or ⑥b and/or ⑦b become

$\bar{D}(\bar{r},t) = \epsilon_0\epsilon_r\,\bar{E}(\bar{r},t)$; $\quad \left(\bar{D} = \epsilon\bar{E}\right.$

$\bar{B}(\bar{r},t) = \mu_0\mu_r\,\bar{H}(\bar{r},t)$ $\quad \bar{B} = \mu\bar{H}$

$\bar{J}(\bar{r},t) = \sigma\,\bar{E}(\bar{r},t)$ $\quad \left.\bar{J} = \sigma\bar{E}\right)$

Fundamentals of EMI
- Electromagnetic fields

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Integral form of Maxwell's equations

- describe EM field properties over regions, action at a distance
- obtained by integrating differential equations, using fundamental properties of vector fields → Gauss' Law, Faraday's Law

① and ⓐ give $\oint_S \bar{D}\cdot d\bar{S} = \int_V \nabla\cdot\bar{D}\, dV = \int_V \rho\, dV = Q$

(divergence thm.) (Maxwell.)

①a $\Psi = \oint_S \bar{D}\cdot d\bar{S} = Q$ Gauss' Law

This says the total electric flux leaving a closed surface equals the enclosed charge

Similarly ② and ⑨ give ②a $\oint_S \bar{B}\cdot d\bar{S} = 0 = Q_m$

ie. the total magnetic flux leaving a closed surface is zero OR magnetic flux always links with itself

Gauss' Law $\Psi = \oiint_S \bar{D}\cdot d\bar{S} = \iiint_V \rho\, dV = Q$ (charge = flux)

by divergence theorem $\oiint_S \bar{D}\cdot d\bar{S} = \iiint_V \nabla\cdot\bar{D}\, dV$

equating terms inside the volume integrals... $\boxed{\nabla\cdot\bar{D} = \rho}$ ①

Similarly, for magnetic fields, $\oiint_S \bar{B}\cdot d\bar{S} = 0$, ∴ $\boxed{\nabla\cdot\bar{B} = 0}$ ②

③ and ⑩ give $\text{emf} = \oint_l \bar{E}\cdot d\bar{l} = \int_S \nabla\times\bar{E}\cdot d\bar{S} = -\frac{\partial}{\partial t}\int_S \bar{B}\cdot d\bar{S} = -\frac{\partial\Phi}{\partial t}$

(Stokes thm.) (Maxwell)

③a $\text{emf} = -\frac{\partial\Phi}{\partial t}$ Faraday Law

This says that the voltage around a loop l equals (minus) the rate of change of magnetic flux linking l.

Faraday's Law $V = \oint \bar{E}\cdot d\bar{l} = -\frac{\partial}{\partial t}\iint_S \bar{B}\cdot d\bar{S} = -\frac{\partial\Phi}{\partial t}$

by Stokes theorem $\oint \bar{E}\cdot d\bar{l} = \iint_S \nabla\times\bar{E}\cdot d\bar{S}$

equating terms inside the surface integrals... $\boxed{\nabla\times\bar{E} = -\frac{\partial\bar{B}}{\partial t}}$ ③

Similarly, for Ampere's Law ...... $\boxed{\nabla\times\bar{H} = \bar{J} + \frac{\partial\bar{D}}{\partial t}}$ ④

∇ = gradient operator (is "slope" in n-dimensional space of a scalar field)
in 3D rectilinear coordinates $\nabla = \mathbf{i}\frac{\partial}{\partial x} + \mathbf{j}\frac{\partial}{\partial y} + \mathbf{k}\frac{\partial}{\partial z}$

Fundamentals of EMI
- Electromagnetic fields

Integral form of Maxwell's equations

describe EM field properties over regions, instead of at points

→ Ampere's Law

④ and ⑩ give

$$mmf = \underbrace{\int_l \bar{H}\cdot d\bar{l} = \int_s \nabla\times\bar{H}\cdot d\bar{s}}_{\text{Stokes}} = \underbrace{\int_s \left(\bar{J} + \frac{\partial \bar{D}}{\partial t}\right)\cdot d\bar{s}}_{\text{Maxwell}} = \left(\sigma + \epsilon\frac{\partial}{\partial t}\right)\int_s \bar{E}\cdot d\bar{s}$$

$$= I_c + \frac{\partial \Psi}{\partial t}$$

$$= I_c + I_D = I_{tot}$$

④a $mmf = \oint_l \bar{H}\cdot d\bar{l} = I_{tot}$ Amperes Law.

This says that the magnetomotive force around a loop l equals the current linking l. Maxwell modified Ampere's law by adding a "displacement" component of current ($\bar{J}_D = \frac{\partial \bar{D}}{\partial t}$) so Ampere's law obeyed charge continuity ($\nabla\cdot\bar{J} = -\frac{\partial \rho}{\partial t}$) for time-varying fields (before Maxwell, Ampere's law only obeyed charge continuity for static fields).
The hypothesis of a displacement current led Maxwell to the prediction of electro-magnetic waves. (The special theory of relativity also shows the existence of E-M waves, but Maxwell worked 50 years before Einstein)

More symbols:

ϵ $[F m^{-1}]$ Electric permittivity
μ $[H m^{-1}]$ Magnetic permeability
σ $[\mho m^{-1}]$ Electric conductivity
Φ [Wb] Magnetic flux
Ψ [C] Electric flux
I_c [A] Conduction current due to flow of electric charge
I_D [A] Displacement current due to change in electric flux.
Q [C] Electric charge
Q_m [Wb] Magnetic charge (non existent, by ②a)
$\bar{x}, \bar{y}, \bar{z}$ unit vectors in x, y and z directions.

(Note: the dependence of quantities on time and or space is often omitted for brevity, don't let this confuse you!)

Fundamentals of EMI
- Sinusoidal electromagnetic waves

Let medium be linear, homogeneous + isotropic

Then

$$\nabla \times \bar{E} = -\frac{\partial \bar{B}}{\partial t} = -\mu \frac{\partial \bar{H}}{\partial t}$$

$$\nabla \times \bar{H} = \bar{J} + \frac{\partial \bar{D}}{\partial t} = \sigma \bar{E} + \epsilon \frac{\partial \bar{E}}{\partial t}$$

$$\nabla \times \nabla \times \bar{E} = -\mu \frac{\partial}{\partial t}(\nabla \times \bar{H}) = -\mu\sigma \frac{\partial \bar{E}}{\partial t} - \mu\epsilon \frac{\partial^2 \bar{E}}{\partial t}$$

$$\nabla(\nabla \cdot \bar{E}) - \nabla^2 \bar{E} = -\mu\sigma \frac{\partial \bar{E}}{\partial t} - \mu\epsilon \frac{\partial^2 \bar{E}}{\partial t}$$

$$\therefore \quad \nabla^2 \bar{E} - \mu\sigma \frac{\partial \bar{E}}{\partial t} - \mu\epsilon \frac{\partial^2 \bar{E}}{\partial t^2} = \nabla \rho/\epsilon = 0 \qquad (1)^*$$

* Let $\rho = 0$ (eg. in free space, homogeneous conductors)

then in 3D rectilinear coordinates (1) is

$$\frac{\partial^2 E_x}{\partial x^2} + \frac{\partial^2 E_x}{\partial y^2} + \frac{\partial^2 E_x}{\partial z^2} = \mu\sigma \frac{\partial E_x}{\partial t} + \mu\epsilon \frac{\partial^2 E_x}{\partial t^2}$$

$$\frac{\partial^2 E_y}{\partial x^2} + \frac{\partial^2 E_y}{\partial y^2} + \frac{\partial^2 E_y}{\partial z^2} = \mu\sigma \frac{\partial E_y}{\partial t} + \mu\epsilon \frac{\partial^2 E_y}{\partial t^2} \qquad (1)a$$

$$\frac{\partial^2 E_z}{\partial x^2} + \frac{\partial E_z}{\partial y^2} + \frac{\partial E_z}{\partial z^2} = \mu\sigma \frac{\partial E_z}{\partial t} + \mu\epsilon \frac{\partial^2 E_z}{\partial t^2}$$

Similar relations hold for the components of $\bar{H}$

*NB: $\rho = 0$ was assumed implicitly in using Ohm's Law, $\bar{J} = \sigma \bar{E}$, rather than charge continuity, $\nabla \cdot \bar{J} = -\partial\rho/\partial t$, in deriving eqn 2.

Proof: $\underbrace{\nabla \cdot \nabla \times \bar{H} = 0}_{\text{by vector identity}} = \underbrace{\nabla \cdot (\sigma + \epsilon\, \partial/\partial t)\bar{E}}_{\text{Maxwell + Ohm}} = (\sigma + \epsilon\, \partial/\partial t)\frac{\nabla \cdot \bar{D}}{\epsilon} = (\sigma + \epsilon\, \partial/\partial t) \cdot \rho/\epsilon$

$\therefore \rho = 0$ (or static field in perfect insulator)

Assuming the fields vary sinusoidally* in time, the Helmholtz wave eqns become (time-harmonic form)

$$\nabla^2 \hat{E}(\bar{r}) = +j\omega\mu(\sigma + j\omega\epsilon)\,\hat{E}(\bar{r}) \qquad (1)a$$

$$\nabla^2 \hat{H}(\bar{r}) = +j\omega\mu(\sigma + j\omega\epsilon)\,\hat{H}(\bar{r}) \qquad (2)a$$

*(NB in linear media, E-M waves with non-sinusoidal and even non-periodic time-variation can be synthesised by superposition of waves with appropriate frequencies + amplitudes)

Also, for simplicity, let

1) $\hat{E}(\bar{r}) = \bar{x} E_x(\bar{r})$ (ie the (electric) field is linearly polarised in the x-direction, so $E_y = E_z = 0$) (other polarisations (eg circular) can be obtained by superposition)

then since $\nabla \times \hat{E} = -j\omega\mu\hat{H}$

$$\bar{y}\frac{\partial E_x}{\partial z} - \bar{z}\frac{\partial E_x}{\partial y} = -j\omega\mu\hat{H} \qquad (H_x = 0)$$

and let

2) $\hat{H}(\bar{r}) = \bar{y} H_y(\bar{r})$ (ie $H_z = 0$ also, this leads to one TEM wave travelling in the z-direction)

from above $\frac{\partial E_x}{\partial y} = 0$, and from $\nabla \cdot \hat{E} = 0$, $\frac{\partial E_x}{\partial x} = 0$

also from $\nabla \times \hat{H} = (\sigma + j\omega\epsilon)\hat{E}$, $\frac{\partial H_y}{\partial x} = 0$, and from $\nabla \cdot \hat{B} = 0$, $\frac{\partial H_y}{\partial y} = 0$

$$= \bar{z}\frac{\partial H_y}{\partial x} - \bar{x}\frac{\partial H_y}{\partial z}$$

$\therefore$ E_x and H_y only vary in the z-direction

Under assumptions (1) and (2) above, the wave equations (1)a, (2)a become

$$\frac{\partial^2 E_x}{\partial z^2} - \gamma^2 E_x = 0 \quad , \quad \frac{\partial^2 H_y}{\partial z^2} - \gamma^2 H_y = 0 \qquad (1),(2)\,b$$

where $\gamma^2 = j\omega\mu(\sigma + j\omega\epsilon)$ (γ is called the propagation constant)

Fundamentals of EMI
- Sinusoidal electromagnetic waves

By the usual methods, or by inspection, the solution to equation ①b is,

$$E_x = E_x^+ e^{-\gamma z} + E_x^- e^{+\gamma z}$$

so $\bar{E} = \bar{x}\,Re\{E_x^+ e^{j\omega(t+j\frac{\gamma}{\omega}z)} + E_x^- e^{j\omega(t-j\frac{\gamma}{\omega}z)}\}$

$= Re\{\hat{E}(z)e^{j\omega t}\}$

which you should recognise as the sum of a forwards and backwards travelling wave. (since γ complex, in general, the wave may be attenuating as it propagates).

Once $\bar{E}$ is known, $\bar{H}$ follows by (or vice versa)

$$\hat{H} = j\frac{1}{\omega\mu}(\nabla \times \hat{E})$$

$\therefore$ $H_y = -j\frac{\gamma}{\omega\mu}(E_x^+ e^{-\gamma z} - E_x^- e^{+\gamma z})$ where $\frac{E_x^+}{H_y^+} = -\frac{E_x^-}{H_y^-} = j\frac{\omega\mu}{\gamma} = \eta$

$= H_y^+ e^{-\gamma z} + H_y^- e^{+\gamma z}$

η is the INTRINSIC IMPEDANCE of the medium

So $\bar{H} = \bar{y}\,Re\{\frac{E_x^+}{\eta} e^{j\omega(t+j\gamma/\omega z)} - \frac{E_x^-}{\eta} e^{j\omega(t-j\gamma/\omega z)}\}$

It has thus been shown that a possible solution of the Helmholtz equations is a wave, travelling in a linear, homogeneous and isotropic medium without boundaries such that the E field and H field are everywhere perpendicular to each other and to the direction of propagation (given by $\bar{E}\times\bar{H}$).

Important parameters of the medium and wave:

$\gamma = \alpha + j\beta$ propagation constant of the medium.

$\alpha = Re(\gamma)$ attenuation constant of the medium [Np.m⁻¹]

$\beta = Im(\gamma)$ phase constant of the medium [rad.m⁻¹]

$\eta = j\omega\mu/\gamma$ intrinsic impedance of the medium [Ω]

$\omega/\beta = c$ velocity of wave in the medium [ms⁻¹]

$\lambda = 2\pi/\beta$ wavelength of wave in the medium [m]

$n = \frac{c_0}{c}$ refractive index of the medium [1]

1) In free space $\sigma = 0$, $\mu = \mu_0$, $\epsilon = \epsilon_0$

$\therefore$ $\gamma = j\omega\sqrt{\mu_0\epsilon_0}$ $(\beta = \omega\sqrt{\mu_0\epsilon_0})$

and the forward travelling EM wave is $\hat{E} = \bar{x}.E_x^+.e^{-j\sqrt{\mu_0\epsilon_0}\,z}$,

NB: $\bar{E}, \bar{H}$ in phase if η real

$\bar{E} = \bar{x}\,E_x^+ \cos[\omega(t - \sqrt{\mu_0\epsilon_0}\,z)] = \bar{x}E_x^+\cos(\omega t - \beta z)$

$\bar{H} = \bar{y}\,E_x^+/\eta \cos[\omega(t - \sqrt{\mu_0\epsilon_0}\,z)] = \bar{y}\frac{E_x^+}{\eta}\cos(\omega t - \beta z)$

In free space, EM waves travel at

$$c_0 = \frac{1}{\sqrt{\mu_0\epsilon_0}} = 3\times10^8\ [ms^{-1}]$$

and the intrinsic impedance of free space is $\eta = \sqrt{\frac{\mu_0}{\epsilon_0}} = 377[\Omega]$

2) In good conductors, conduction current dominates over displacement current so $\sigma \gg \omega\epsilon$

$\therefore$ $\gamma \doteq \sqrt{j\omega\mu\sigma} = \sqrt{\frac{\omega\mu\sigma}{2}}(1+j) = \alpha + j\beta$

and the forward travelling $\bar{E}$ wave is ($\bar{H}$ similar but with phase shift)

$\bar{E} = \bar{x}\,E_x^+ \exp(-\sqrt{\frac{\omega\mu\sigma}{2}}\,z)\cos[\omega(t - \sqrt{\frac{\mu\sigma}{2\omega}}\,z)]$

$= \bar{x}\,E_x^+ e^{-\alpha z}\cos(\omega t - \beta z)$

The phase velocity is $c = \sqrt{\frac{2\omega}{\mu\sigma}}$ [ms⁻¹]

and intrinsic impedance $\eta = \frac{j\omega\mu}{\gamma} = (1+j)\sqrt{\frac{\omega\mu}{2\sigma}}$ [Ω]

The skin depth is a measure of depth to which the E-M wave travels into the conductor

$$\delta = \frac{1}{\alpha} = \sqrt{\frac{2}{\omega\mu\sigma}}\ [m]$$

$\left(E_0 e^{-\alpha\delta} = E(\delta) = \frac{E_0}{e}\right)$

Fundamentals of EMI
- Boundary conditions

WHEN A WAVE ENCOUNTERS A BOUNDARY BETWEEN MEDIA WITH DIFFERENT INTRINSIC IMPEDANCES, THEN IN ORDER FOR THE FIELDS AT THE INTERFACE TO SATISFY BOUNDARY CONDITIONS, A REFLECTED WAVE AND TRANSMITTED WAVE MAY RESULT

(ANALOGIES: 1) REFLECTIONS ON TRANSMISSION LINES FROM UNMATCHED SOURCES OR LOADS $Z_L \neq Z_0$, $Z_S \neq Z_0$

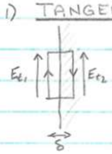
2) NON-MAXIMUM POWER TRANSFER IN CIRCUITS WITH LOAD UNMATCHED TO SOURCE $Z_L \neq Z_S^*$

HOWEVER, THESE ANALOGIES ARE ONLY GOOD WHEN CONSIDERING PLANE WAVES NORMALLY INCIDENT AT PLANE INTERFACES. WAVES CAN ALSO MEET INTERFACES OBLIQUELY).

REVIEW OF BOUNDARY CONDITIONS

Basically tangential components of $\bar{E}$ and $\bar{H}$ (or, alternatively, normal components of $\bar{D}$ and $\bar{B}$) must be continuous across the boundary.

1) **TANGENTIAL $\bar{E}$ DOES NOT CHANGE AT BOUNDARIES**



$$\oint_L \bar{E}.d\bar{l} = \frac{\partial}{\partial t}\int_S \bar{B}.d\bar{s} \Longrightarrow 0 \quad \text{as } \delta \to 0$$

$$\therefore E_{t1}\Delta l - E_{t2}\Delta l = 0$$

$$\therefore E_{t1} = E_{t2} \qquad [\text{ie. } \bar{n}\times(\bar{E}_1 - \bar{E}_2) = 0]$$

NB if one of the media is a perfect conductor, then $E_{t1} = E_{t2} = 0$ [ie. $\bar{n}\times\bar{E} = 0$]

2) **TANGENTIAL $\bar{H}$ DOES NOT CHANGE AT BOUNDARIES** (unless there is a surface current)

Using same setup as in (1),

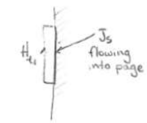
$$\oint_L \bar{H}\cdot d\bar{l} = \oint_S \left(\bar{J} + \frac{\partial \bar{D}}{\partial t}\right).d\bar{s} \Rightarrow 0 \quad \text{as } \delta \to 0 \quad \text{if } \lim_{\delta\to 0} J, D = 0 \text{ (ie } J, D \text{ finite)}$$

$$\therefore H_{t1} = H_{t2} \qquad [\text{ie. } \bar{n}\times(\bar{H}_1 - \bar{H}_2) = 0]$$

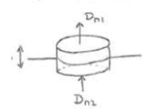
However, in a perfect conductor (eg superconductor) all internal magnetic fields are zero due to surface currents. In that case $\lim_{\delta\to 0} J = J_s$ [A m⁻¹]

if $H_{t2} = 0$, $H_{t1} = J_s$

[ie $\bar{n}\times\bar{H}_1 = \bar{J}_s$]



3) **NORMAL $\bar{D}$ DOES NOT CHANGE AT BOUNDARIES** (unless there is a surface charge)



$$\oint_S \bar{D}.d\bar{s} = \int_V \rho dv \Rightarrow 0 \quad \text{as } \delta\to 0 \quad \text{if } \rho \text{ finite } (\lim_{\delta\to 0}\rho = 0)$$

$$D_{n1}ds_1 - D_{n2}ds_2 = 0$$

$$D_{n1} = D_{n2} \qquad [\text{ie } \bar{n}.(\bar{D}_1 - \bar{D}_2) = 0]$$

However, in a perfect conductor charge resides only at the surface ($\lim_{\delta\to 0}\rho = \rho_s$ [C m⁻²])

then $D_{n2} = 0$, $D_{n1} = \rho_s$

[ie $\bar{n}.\bar{D}_1 = \rho_s$]

4) **NORMAL $\bar{B}$ DOES NOT CHANGE AT BOUNDARIES**

Using same setup as (3)

$$\oint_S \bar{B}.d\bar{s} = 0$$

$$\therefore B_{n1} = B_{n2} \qquad [\text{ie. } \bar{n}.(\bar{B}_1 - \bar{B}_2) = 0]$$

In a superconductor, $B_{n1} = B_{n2} = 0$ [ie. $\bar{n}.\bar{B} = 0$]

SUMMARY OF BOUNDARY CONDITIONS

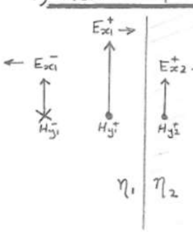
	σ = finite		$\sigma = \infty$	
BC1	$\bar{n}\times(\bar{E}_1 - \bar{E}_2) = 0$	(tangential E)	$\bar{n}\times\bar{E} = 0$	
BC2	$\bar{n}\cdot(\bar{B}_1 - \bar{B}_2) = 0$	(normal B)	$\bar{n}\cdot\bar{B} = 0$	
BC3	$\bar{n}\times(\bar{H}_1 - \bar{H}_2) = 0$, or J_s if $J\to\infty$		$\bar{n}\times\bar{H}_1 = \bar{J}_s$	[A m⁻¹]
BC4	$\bar{n}\cdot(\bar{D}_1 - \bar{D}_2) = 0$, or ρ_s if $\rho\to\infty$		$\bar{n}\cdot\bar{D}_1 = \rho_s$	[C m⁻²]

Fundamentals of EMI
- Reflections of plane waves

➢ Fundamentals of shielding against EMI in far field

- Assumption: plane wave normally incident on plane boundary between two semi-infinite regions (TEM incidence)
- Analogous to reflections of waves on transmission lines (also TEM)

1) NORMALLY INCIDENT

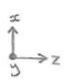


in medium ① (see lecture 2)

$$\hat{E}_1 = \left(E_{x10}^{+} e^{-\gamma_1 z} + E_{x10}^{-} e^{\gamma_1 z}\right)\bar{x}$$
$$\hat{H}_1 = \left(E_{x10}^{+} e^{-\gamma_1 z} - E_{x10}^{-} e^{\gamma_1 z}\right)\frac{1}{\eta_1}\bar{y}$$

in medium ②

$$\hat{E}_2 = \left(E_{x20}^{+} e^{-\gamma_2 z}\right)\bar{x}$$
$$\hat{H}_2 = \left(E_{x20}^{+} e^{-\gamma_2 z}\right)\frac{1}{\eta_2}\bar{y}$$



at the boundary (z=0)

for $\bar{E}$ fields $\quad E_{x10}^{+} + E_{x10}^{-} = E_{x20}^{+}$ $\quad$ by (BC1) $\quad$ ①

for $\bar{H}$ fields $\quad \dfrac{E_{x10}^{+} - E_{x10}^{-}}{\eta_1} = \dfrac{E_{x20}^{+}}{\eta_2}$ $\quad$ by (BC3) $\quad$ ②

in order to define E_{x10}^{-}, E_{x20}^{+} $(H_{y10}^{-}, H_{y20}^{+})$ in terms of the incident E_{x10}^{+} (H_{y10}^{+}) we define

reflection coefficient $\quad \rho_o = \dfrac{E_{x10}^{-}}{E_{x10}^{+}}$

transmission coefficient $\quad \tau_o = \dfrac{E_{x20}^{+}}{E_{x10}^{+}}$

then ① and ② may be written:

① $\quad 1 + \rho_o = \tau_o$

② $\quad 1 - \rho_o = \tau_o \eta_1 / \eta_2$

$$\therefore \quad \rho_o = \frac{\eta_2 - \eta_1}{\eta_2 + \eta_1} \qquad \tau_o = \frac{2\eta_2}{\eta_2 + \eta_1}$$

The fields on either side of the boundary are therefore

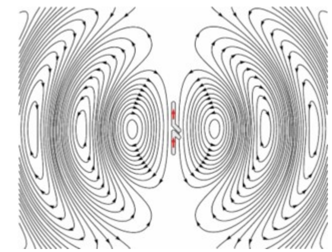
③ in medium ① $\quad \hat{E}_1 = \bar{x} E_{x10}^{+}\left[e^{-\gamma_1 z} + \underbrace{\left(\frac{\eta_2 - \eta_1}{\eta_2 + \eta_1}\right)}_{\rho_o} e^{\gamma_1 z}\right]$, $\quad \hat{H}_1 = \bar{y}\frac{E_{x10}^{+}}{\eta_1}\left[e^{-\gamma_1 z} - \left(\frac{\eta_2 - \eta_1}{\eta_2 + \eta_1}\right)e^{\gamma_1 z}\right]$

④ in medium ② $\quad \hat{E}_2 = \bar{x} E_{x10}^{+} \underbrace{\frac{2\eta_2}{\eta_2 + \eta_1}}_{\tau_o} \cdot e^{-\gamma_2 z}$, $\quad \hat{H}_2 = \bar{z} \times \hat{E}_2 / \eta_2$

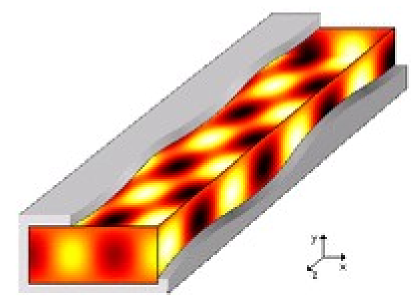
The fraction of incident power at the interface $\begin{cases} \text{transmitted} & P_2 / P_1^{+} = 1 - |\rho_o|^2 \quad ⑤ \\ \text{reflected} & P_1^{-} / P_1^{+} = |\rho_o|^2 \quad ⑥ \end{cases}$

Fundamentals of EMI - EM waves - summary

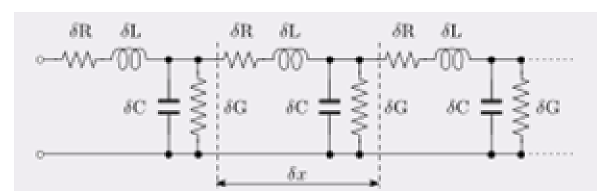
- **EM waves in free space (EM radiation)**
  - antenna = V←→E, I←→ H "transducer"



- **Guided EM waves**
  - TEM (two conductors → transmission line; V, I)
  - TE, TM (one conductor, dielectric → waveguide; E,H)



- **For all waves, two fundamental parameters:**
  - Wave impedance, $Z(z) = E_x(z)/H_y(z) = V(z)/I(z)$
    - real part → energy dissipation
    - imaginary part → energy storage
  - Propagation constant, $\gamma = \alpha + j\beta = j\omega\mu.(\sigma + j\omega\varepsilon)^{1/2} = j\omega\mu/Z$
    - α → change in amplitude with distance – loss, evanescent fields
    - β → change in phase with distance – propagating fields
  - $Z = j\,\omega\,\mu\,/\,\gamma$



Fundamentals of EMI
- EM wave impedance, near vs far field

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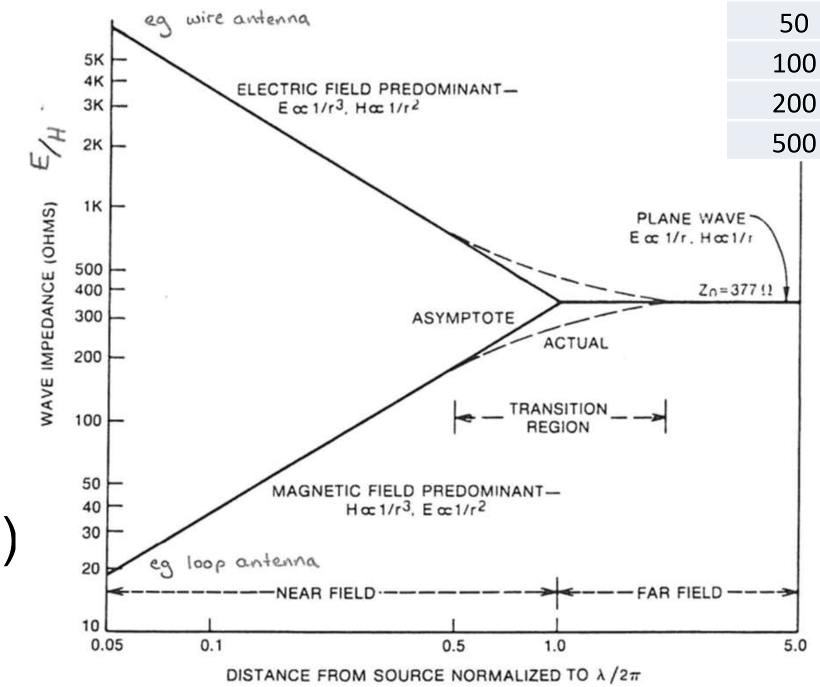
EM impedance depends on
- type of source (E, H)
- distance from source
- frequency

EM coupling depends on impedance matching
- Conducted
  - Common impedance
  - Guided-wave
- Near-field (induced)
  - Capacitive (E-field)
  - Inductive (H-field)
- Far-field (radiated, free space)
  - $E/H = Z_0 = 377\ \Omega$

➢ **Most low frequency EMI conducted or induced**

$D_{near\text{-}field} < \lambda/2\pi < D_{far\text{-}field}$

f [MHz]	$\lambda/2\pi$ [m]
1	47.75
2	23.87
5	9.55
10	4.77
20	2.39
50	0.95
100	0.48
200	0.24
500	0.10



Adapted from Ott, Noise reduction techniques in electronic systems, Wiley 1976.

PART 3 – Overview
- EMI reduction strategies

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1. Reduce EMI generation

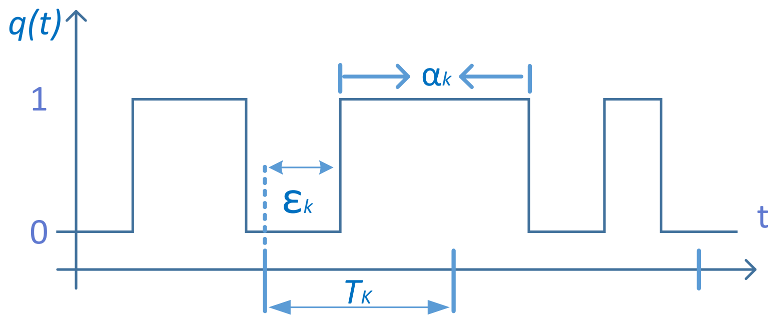
- Modulate PWM parameter(s) – e.g. switch phase (or frequency)
  - spreads EMI over more frequencies…
  - …reduces peak power spectral spectral density (no change to average PSD)
- "Soft" switching of active devices – generate fewer harmonics
  - resonant converters - limit number of harmonics
  - slow switching - reduce di/dt, dv/dt - limit bandwidth of harmonics
  - additional circuitry, complexity, loss

2. Reduce EMI coupling, impede energy transfer

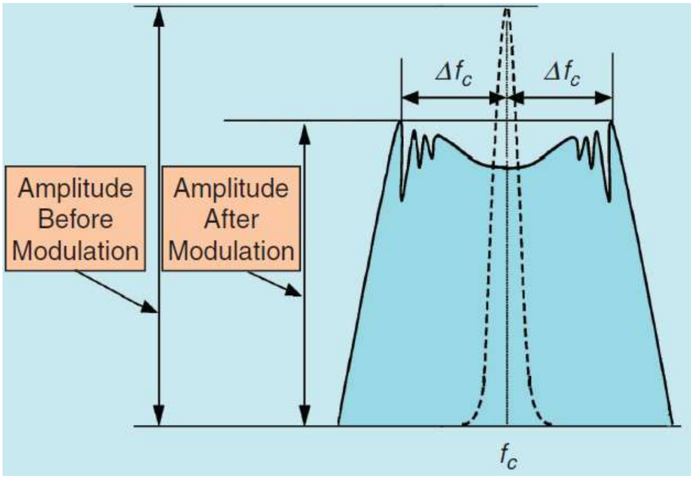
- Circuit layout (conducted, induced)
- Filtering (conducted, induced)
  - cost & size
- Shielding (induced, radiated EMI)
  - weight + cost
  - often need to allow for ventilation

Reduce EMI Generation
- PWM techniques

Modulating one or more parameters of a switch driving signal, q(t)



redistributes energy in frequency domain



Reduce EMI Generation
- Classification of PWM techniques

MACQUARIE University

Modulation Style	SCHEME	Sub classification	Pulse modulation parameters				
			Switching cycle duration (T_k)	Pulse width (a_k)	Pulse position (ε_k)	Duty ratio (a_k/ T_k)
Periodic OR Aperiodic (pseudo-random, chaotic, deterministic) OR Random*	PWM	--	Fixed	Variable	Fixed	Variable
	PPM	--	Fixed	Fixed	Variable	Fixed
	DRM+PPM+ fixed carrier frequency		Fixed	Variable	Variable	Variable
	CFM	CFM with Fixed Duty cyle	Variable	Synchronised	Fixed	Fixed
		CFM with varying Duty cyle	Variable	Fixed or Variable (not synched)	Fixed	Variable
	CFM+PPM+ fixed duty ratio	--	Variable	Variable (synched)	Variable	Fixed (synched)
	CFM+PPM+DRM	--	Variable	Variable	Variable	Variable

PWM= Pulse Width Modulation, PPM= Pulse Position Modulation, DRM= Duty Ratio Modulation, CFM= Carrier Frequency Modulation

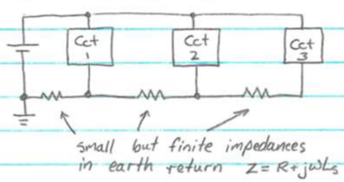
Reduce EMI Generation
- PWM techniques

- **Additional material to be included**
  - Example of PWM technique for EMI mitigation
  - “Soft” Switching

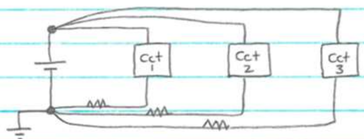
EMI Reduction
- Minimising common impedance coupling

③ Common impedance coupling.

This doesn't involve coupling of electric or magnetic fields via non-conductive media, but coupling between circuits which have some conductive part in common. The most common such situation arises in "grounding" different parts of a circuit.



Currents through circuit 3 to earth raise the "earth" potential seen by ccts 1 and 2. Real problems occur if, for example, circuit 1 is senstive analog ccty, and 2 or 3 are digital circuits.



preferred connection for circuits that require common "ground". (at "low" eg audio - frequencies)

Currents in any one circuit do not affect the "earth" reference of any other.

The above situation is one of the simplest that can occur. Many other more complex situations arise, but won't be covered here because its not really "fields + waves" material. Refer to the book by Ott if interested.

Near Field EMI Reduction
- Minimising capacitive coupling

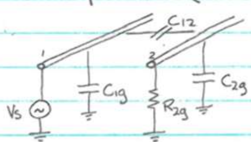
Ⓐ NEAR FIELD

In this case we can consider the effect of electric and magnetic fields separately. In fact, circuit theory can be used to model the electric and magnetic interactions involved.

The most common interference problems encountered (especially within a piece of equipment) are due to near field coupling. eg at 1MHz, the near field extends 50m from the source.

PROBLEMS and SOLUTIONS.

① Capacitive (electric near-field) coupling.

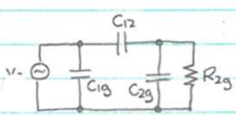


two conductors above a ground plane.

The capacitance between two wires (or any other shaped conductors) can be calculated using methods learnt in "electrostatics".

eg for two parallel wires $C \approx \frac{27.8\,\epsilon_r}{\ln(2h/a)}$ [pF/m]

(2h = distance between conductors
a = radius of conductors)



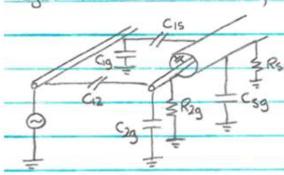
if $R_{2g} \ll Z_{C12} // Z_{C2g}$ then $V_2 = j\omega R C_{12} V_s$

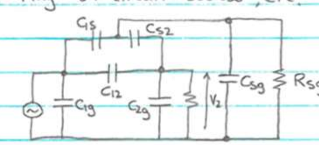
so clearly pickup on conductor 2 is proportional to frequency, impedance to ground, capacitance from source, strength of source

Solutions: i) lower impedance wrt earth of "receiver"
ii) reduce capacitive coupling. – moves wires apart – add shield
iii) increase C_{1g}, C_{2g} – keep wires close to ground plane

Shielding against electric field coupling:

egs. i) coaxial cable, ii) guard ring on circuit boards, etc.

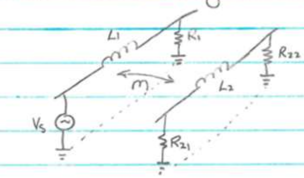




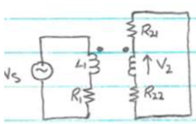
If the shield is grounded*, and conductor 2 is kept totally within the shield* then all electric field coupling between conductors 1 and 2 is eliminated.
(Note:* there is no such thing as a perfect ground or perfect shield in practice, but they can be made pretty good, with care)

Near Field EMI Reduction
- Minimising inductive coupling

② Inductive (Magnetic near field) coupling.



The mutual inductance between two loops carrying low frequency currents can be calculated using methods learned in magnetostatics.



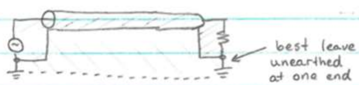
$V_2 = j\omega M I_1$

Solution: reduce $M = \frac{\mu}{4\pi} \oint_{C_1} \oint_{C_2} \frac{d\bar{l}_1 \cdot d\bar{l}_2}{|\bar{r}_1 - \bar{r}_2|}$

ie reduce magnetic flux linkages between the two circuits.

- Increase distance between circuits
- decrease area enclosed by each circuit. < length, breadth

eg * twisted pairs of wire



* use shield for return current path (also reduces coupling via eddy currents)

Note: grounding at both ends can cause problems at audio frequencies because of "earth loops" - noise current can flow in shield, and this couples to the centre conductor.

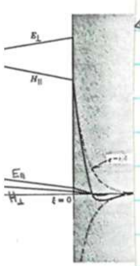
Far Field EMI Reduction
- Minimising radiative coupling

Ⓑ FAR FIELD

In this case the electric and magnetic fields are considered together (ie electromagnetic waves). When a wave encounters an electronic circuit, some of the energy in the wave may be converted to voltages and currents in the circuit which usually are not wanted. (eg when a wave encounters a TV or radio antenna, voltages and currents are induced, except in that case its usually desirable).

The most effective method of reducing radiated interference is shielding. Shielding effectiveness varies with frequency, geometry of shield, direction of incidence and polarisation of the incident wave. We'll consider a plane shield and its effect on a plane wave.

when a plane wave strikes a plane conducting sheet,



i) part is reflected at the first boundary
ii) part is transmitted into the conductor and is attenuated as it travels (ie absorption occurs within the conductor)
iii) part of what remains is reflected at the second boundary (and is further dissipated by conductor)
iv) part is transmitted past the shield.

Total shielding effectiveness is

$$S = A + R + B \quad [dB]$$

where A = absorption loss [dB]
R = reflection loss [dB]
B = correction factor for multiple reflections [dB] in shields less than about one skin depth thick.

Reflection loss

neglecting multiple reflections within the shield, (and absorption)

at first boundary: $E_{ts} = (1+\rho)E_{i1} = \frac{2\eta_s}{\eta_o+\eta_s}E_{i1}$ $\qquad H_{ts} = (1-\rho)H_{i1} = \frac{2\eta_o}{\eta_o+\eta_s}H_{i1}$

at second boundary $E_{t2} = \frac{2\eta_o}{\eta_o+\eta_s}E_{ts} = \frac{4\eta_o\eta_s}{(\eta_o+\eta_s)^2}E_{i1}$ $\qquad H_{t2} = \frac{2\eta_s}{\eta_o+\eta_s}H_{ts} = \frac{4\eta_o\eta_s}{(\eta_o+\eta_s)^2}H_{i1}$

Usually $\eta_o \gg \eta_s$, so
electric field mainly reflected from first boundary
magnetic " " " " second "

and $\frac{E_{t2}}{E_{i1}} = \frac{H_{t2}}{H_{i1}} \doteq \frac{4\eta_s}{\eta_o}$, thus $R = -20\log\left(4\left|\frac{\eta_s}{\eta_o}\right|\right) = 20\log\left(\frac{|Z_o|}{4|Z_s|}\right)$ [dB]
(R = reflection loss)

for conductors $|\eta_s| = \sqrt{\frac{\omega\mu}{\sigma}}$ [Ω], and $\eta_o = 377$ [Ω]

$$\therefore R = 39.5 + 10\log\left(\frac{\sigma}{\omega\mu}\right) \quad [dB]$$

($\sigma_{cu} = 5.82\times10^7$ [℧/m])

Aside:
Note: reflection loss also applies in near field but $Z \neq \eta_o$
for E, $|Z| \doteq 1/2\pi f\epsilon r$
for M, $|Z| \doteq 2\pi f\mu r$
$\therefore R_E$ large
R_M small

Absorption loss

A wave propagating in a conductive medium decreases exponentially in amplitude as $E(z) = E_o e^{-\alpha z} = E_o e^{-z/\delta}$

where $\alpha = \frac{1}{\delta} = \sqrt{\frac{\omega\mu\sigma}{2}}$ [Npm⁻¹]

(Absorption applies equally in near + far fields)

$$\therefore A = 20\log(e^{z/\delta}) = 8.69(d/\delta) \quad [dB]$$

(ie ≈ 9dB per skin depth) (where d = thickness of shield)

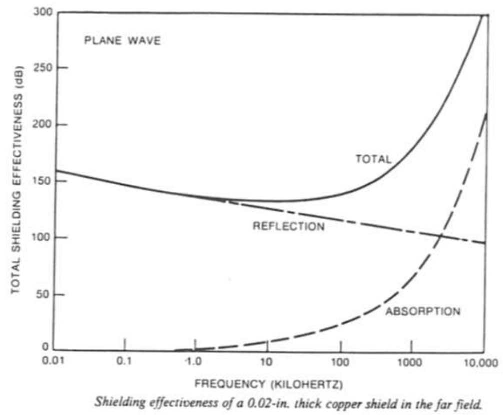
Correction for thin shields

$$B = 20\log(1 - e^{-d/\delta}) \quad [dB]$$

(neglects phase shifts in shield since $d \ll \lambda$)

This correction is usually not required except at low frequencies or in perfect conductors or very thin films.

Far Field EMI Reduction
- Minimising radiative coupling



Shielding effectiveness of a 0.02-in. thick copper shield in the far field.

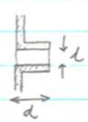
Ott p 151

Note well. Attenuation of waves due to shielding derived above represents the best possible.

* In practice, leakage of waves through seams, joints and holes in the shield is far more significant than leakage through the shield material itself.

- Discontinuities such as holes and seams can form waveguides for the incident radiation
- Leakage occurs even through a waveguide below cutoff:

$$S \doteqdot 30\, d/l \ [dB]$$



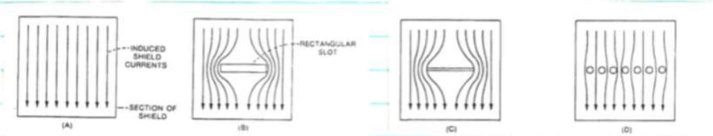
where S = shielding provided by waveguide below cutoff

d = length of guide

l = largest linear dimension

Solutions: - use conductive gaskets at all joints

- keep the linear dimension (not area) of any discontinuities as small as possible.



Effect of shield discontinuity on magnetically induced shield current.

Reduce EMI coupling

- **Additional material to be included**

  Methods – physical (layout), electrical (impedance mismatch)

  - common and differential mode EMI
  - conducted EMI - filtering
  - radiated EMI

    antennas, shielding, Babinet's principle, new composite materials, ventilation

EMI/EMC measurements

- **Additional material to be included**

  EMI/EMC Measurements

  - conducted EMI – LISN
  - time-frequency relationships

Conclusions

- Switch-mode power electronics becoming increasingly common in modern power systems → Internet of Energy.
- Component advances enabling faster switching (ns) at higher frequencies (harmonics to GHz) → efficient, compact converters.

➢ **Broadband EMI noise an increasing problem, especially for low power wireless communications and the Internet of Things.**

- Power spectral density of generated EMI can be reduced using a choice of PWM and switching methods.
- Various EM coupling mechanisms (conducted, induced, radiated)
  - dominant mode of interference depends on frequency, distance from source
  - coupling can be minimized by impedance mismatch (filters, shielding, etc).

➢ **EMI may be minimized by careful design, based on understanding of fundamentals of EMI generation and coupling.**

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