Analog Electronics for Beam Instrumentation

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CERN

June 4-5, 2018

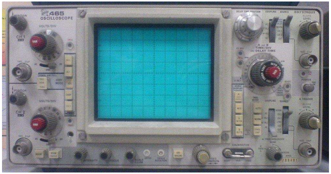
Subjects

- Lab Instrumentation
- Transmission lines
- Transmission line transformers
- Filters
- Noise
- Amplifiers
- EMC
- Radiation effects

Instrumentation

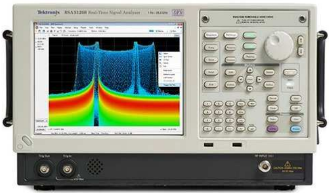
The Oscilloscope

- Plots voltage vs. time
- Maybe the most versatile instrument ever



The Spectrum Analyzer

- Plots signal magnitude vs. frequency
- Good for signal and noise level measurements
- Receiver and mixer diagnostics,
- Distortion measurement
- Chasing interference and stability problems

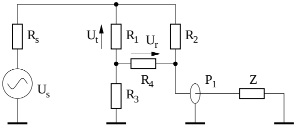


The Network Analyzer

- Frequency-domain analysis of electrical networks
- Measures transmission and reflection vs. frequency
- Complex data format $a + jb$
- Well-defined port impedance, usually 50 Ω
- Usually two ports

The Network Analyzer: 1 port

- Wheatstone bridge
- $R_s = R_1 = R_2 = R_3 = R_4$
- Z : network under test
- $H(f) = \frac{U_r}{U_s} = \frac{Z-R}{8(Z+R)}$

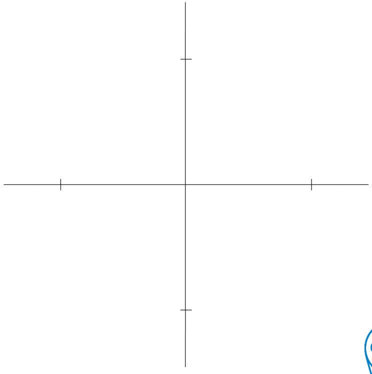


- $H(f)$ is complex
- For all values of Z with real part $>= 0$, $H(f)$ ends up inside a circle of diameter $1/8$

The Network Analyzer: Measuring impedance

- Let's normalize the radius of that circle to unity, so $H = \frac{Z-R}{Z+R}$

- $Z = R$ is in the centre
- $Z \to \infty$ is at $(1, 0)$
- $Z = 0$ sits at $(-1, 0)$
- Z imaginary and positive: Somewhere along the edge of a circle of $r = 1$ above the X-axis
- Z imaginary and negative: Somewhere along the edge of a circle of $r = 1$ below the X-axis



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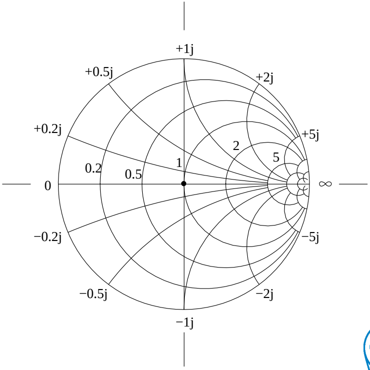
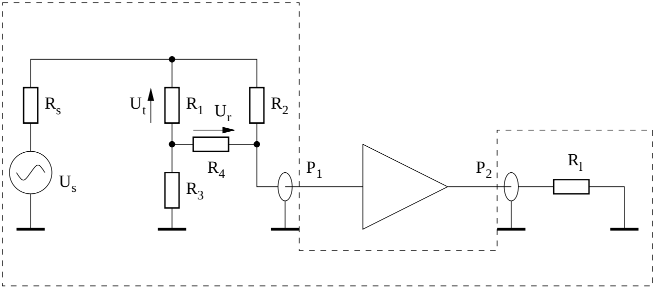


The Network Analyzer: Measuring transmission



- This way the NA can measure the frequency response of amplifiers, filters, etc.
- Often, the two ports are identical and the source U_s can be connected to either

Transmission Lines

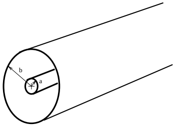
Transmission lines

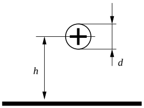
- Confine EM fields between two conductors
- Little radiation loss
- Protected from interference
- Propagation velocity set by material choice
- Wave impedance set by geometry

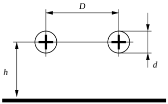
Various geometries

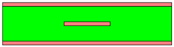
Geometry examples

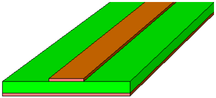
- Coaxial cable
- Wire over ground plane
- Wire pair
- Stripline, Microstrip, Coplanar waveguide





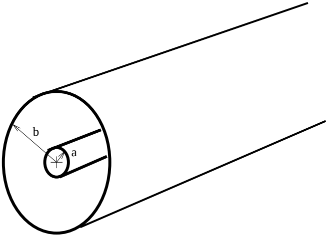






Coaxial cable

- $\mu_0 = 4\pi 10^{-7}$ H/m
- $\varepsilon_0 = \frac{1}{\mu_0 c^2} \approx 8.85$ pF/m
- μ_r Relative magnetic permeability
- ε_r Relative dielectric constant
- $L_0 = \int_a^b \frac{\mu}{2\pi r} dr = \frac{\mu}{2\pi} \ln \frac{b}{a}$
- $C_0 = \frac{1}{\int_a^b \frac{1}{2\pi\varepsilon} dr} = \frac{2\pi\varepsilon}{\ln \frac{b}{a}}$
- $Z_0 = \sqrt{\frac{L_0}{C_0}} \approx 60\sqrt{\frac{\mu_r}{\varepsilon_r}} \ln \frac{b}{a}$
- $v_0 = \frac{1}{L_0 C_0} = \frac{c}{\sqrt{\mu_r \varepsilon_r}}$

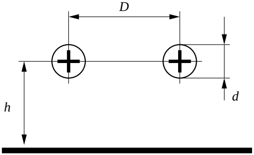


Impedance, Propagation velocity

- We used to have lots of formulae, some closed form, some issued from fits to laborious measurements, to calculate the properties of transmission lines for all sorts of geometries.
- We don't do that anymore.
- These days, we use EM simulation software, like 'atlc' for simple transmission lines, or like e.g. 'HFSS' or 'CST Microwave Studio' for full structure simulation.

$$Z_0 = \frac{69}{\sqrt{\varepsilon_r}} \log\left(\frac{4h}{d}\sqrt{1+\left(\frac{2h}{D}\right)^2}\right)$$

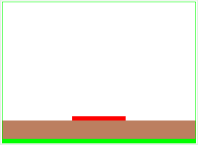
with $d << D$ and $d << h$.
(Common-mode impedance!)

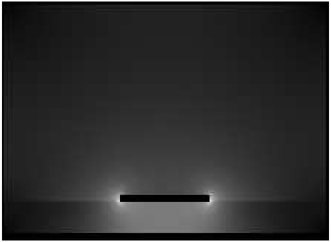


An atlc example

atlc

- Create a picture of the cross-section in BMP format
- atlc strip-atlc.bmp
- strip-atlc.bmp 2 Er= 2.53 Zo= 40.999 Ohms C= 129.5 pF/m L= 217.7 nH/m v= 1.884e+08 m/s vf= 0.628





Coaxial Connectors



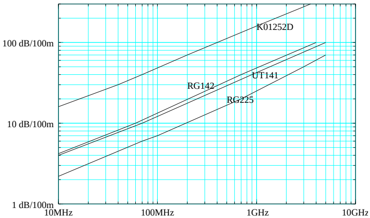


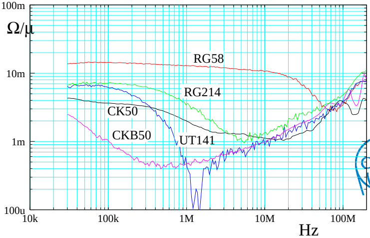

Coaxial connector variants

- Cable connectors, crimp, solder or screw clamp, straight or 90°
- Panel or bulkhead connectors
- Microstrip connectors
- PCB mount connectors
- PCB edge-mount connectors
- 50 Ω or 75 Ω
- etc, etc, etc.

Coaxial Cable Limitations

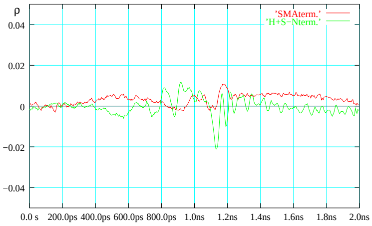
- Losses (below, left)
  Caused by resistance in the conductors and dielectric losses in the insulators. Skin effect makes this worse.
- Screening effectiveness (below, right)
  Screen resistance and density.
- Power handling limits
  Size of the cable, thickness and density of dielectric.

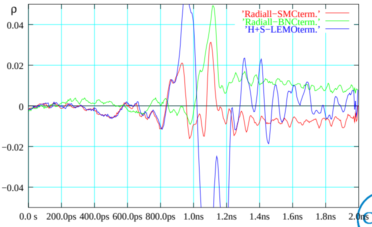




Connector quality

- SMA: Very good. Usable up to 26 GHz.
- N: Rugged and reliable. Usable up to 18 GHz.
- SMC: Very good up to 10 GHz. Tiny and somewhat fragile.
- BNC: Easy to use. Usable up to 4 GHz.
- LEMO: Even easier to use. Usable up to 1.4 GHz.



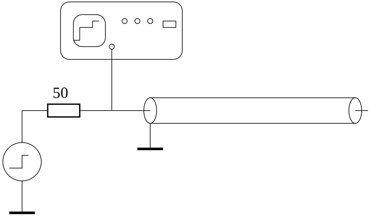


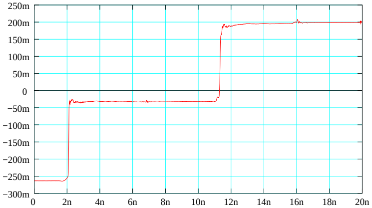
TDR plots of some connector types

Time Domain Reflectometry

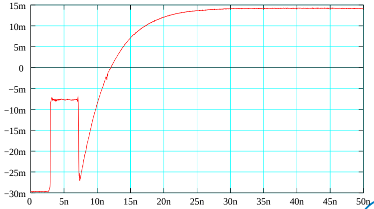
Time Domain Reflectometry

- Launch a fast step into a structure
- Observe reflection
- $\rho = \frac{R - Z_0}{R + Z_0}$



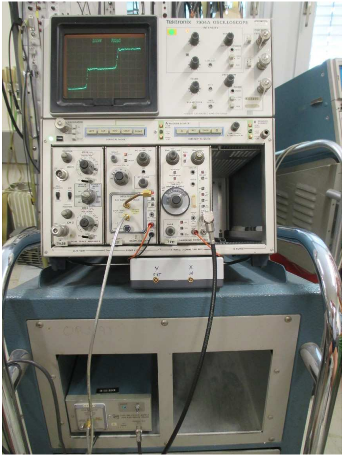


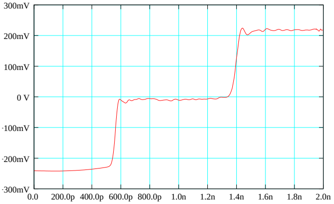
Open Circuit



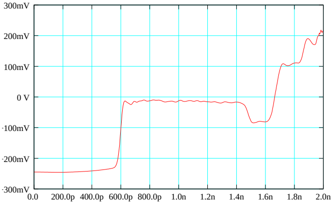
With 82 pF at the end

TDR vintage hardware: Tektronix 7904A with S-52 pulse generator and S-6 sampler



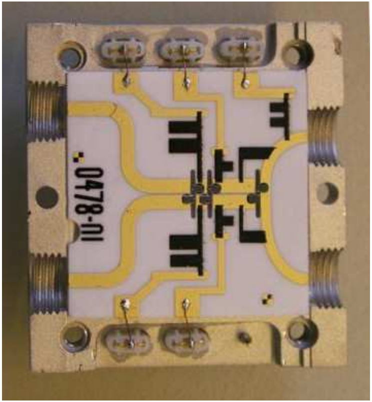


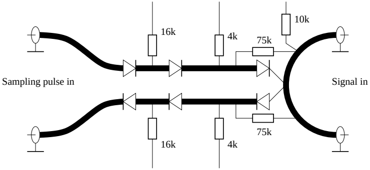
6cm of UT141, open end



6cm of UT141, an SMA T and two SMA M-M adapters

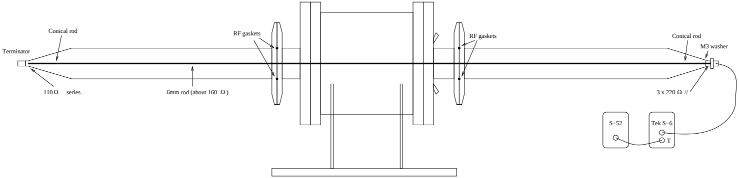
TDR vintage hardware: Tektronix S-6 sampler

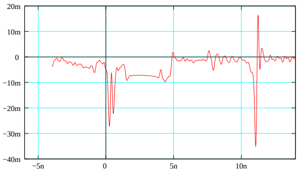




- Risetime: 30 ps (Still respectable!)

TDR application example: Measuring a WCM





- A rod through a beam transformer
- TDR identifies discontinuities

Transmission line transformers

Transmission line transformers

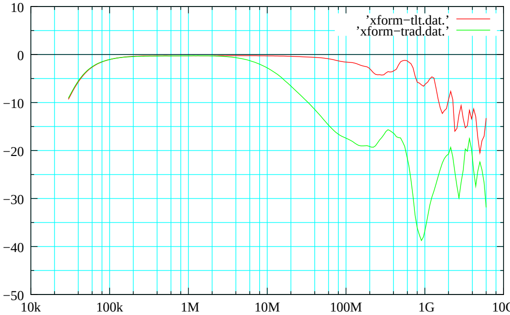
- It's possible to make very good transformers by exploiting transmission line effects

Possible uses:

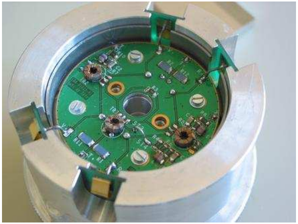
- Scaling voltage, current and impedance
- Impedance matching
- Noise matching
- Combiners and splitters
- Single-ended ⇔ differential conversion
- Feedback elements in low-noise amplifiers
- Hybrids and directional couplers
- ...

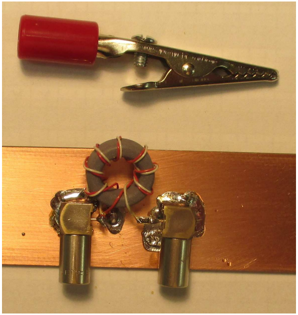
Transmission line transformers

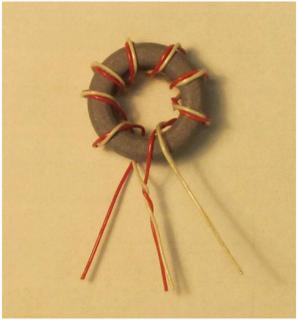
- Compare traditional (T) and transmission line transformers (B)

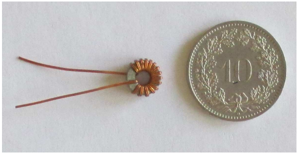


Transmission line transformers: Pictures



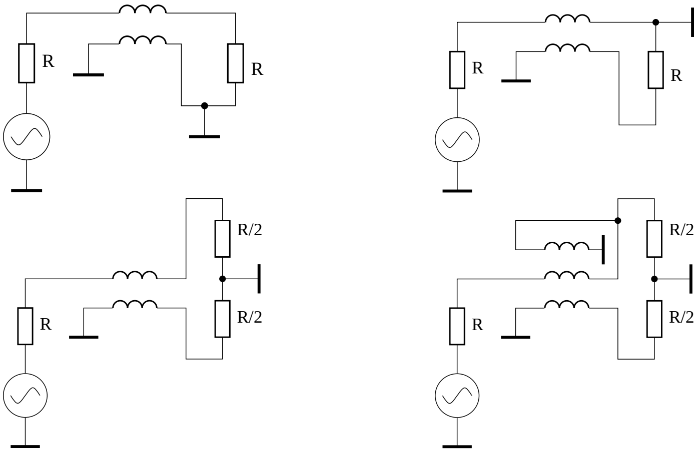






Transmission line transformers

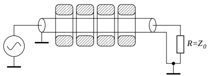
- Wire Baluns

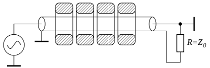


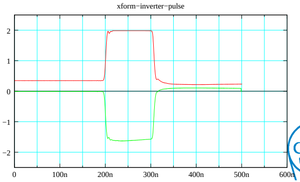
Transmission line transformers

A transmission line balun

- The common mode impedance of an arms-length of coax exceeds the characteristic impedance above a few MHz.
- If you wind the coax on a ferrite toroid, it's easy to bring that down to ≈ 100 kHz without affecting the maximum frequency
- It no longer matters (much) which side you connect to ground!

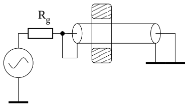


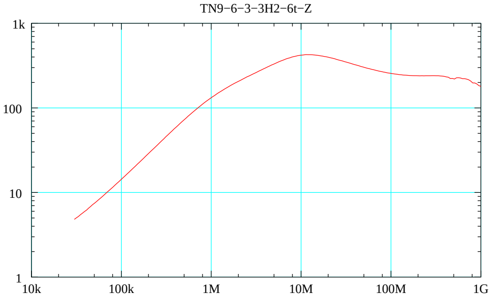




Equivalent circuit for the common mode

- The common-mode impedance of the windings sets the lower cut-off frequency
- This impedance is not a pure inductance, but that doesn't matter if it's significantly higher than the load impedance
- Low loss magnetics are not required





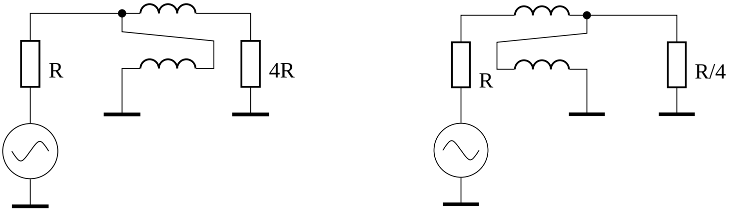
Impedance of a 6-turn coil on a small high-permeability toroid core

Transmission line transformers

- It's customary to specify the *impedance* ratio
- ... which is the square of the voltage ratio
- The transmission line doesn't *have* to be coax
  - Twisted pairs
  - Parallel wires
- The lines may be wound as several turns on a single core
  - ... or a single pass through several cores
  - ... or some combination
- Windings with the same common-mode voltage may share cores
- High μ_r cores extend LF cut-off frequency downward

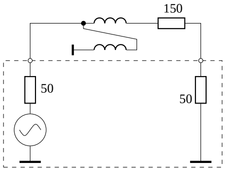
Transmission line transformers

- Wired 4-1 transformers

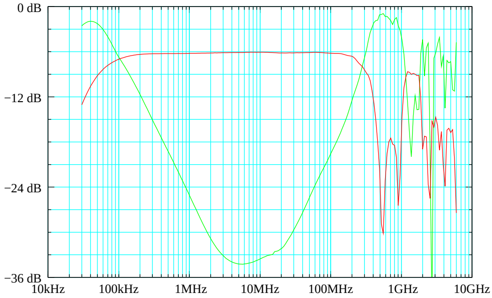


- These transformers have a null where the transmission line length is $\lambda/2$
- The wire length must be short compared to the wavelength at the highest frequency

Transmission line transformers



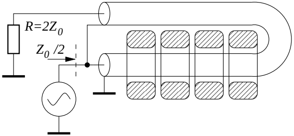
Test circuit for Ruthroff 1:4 transformer

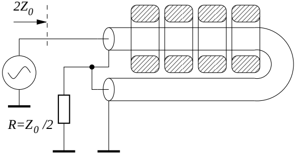


Frequency response of wire-wound Ruthroff 1-4 transformer

Transmission line transformers

- 4-1 transformers with coax

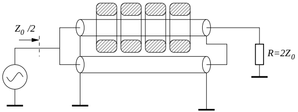


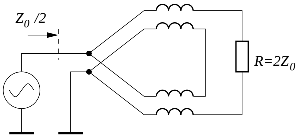


- These transformers have a null where the transmission line length is $\lambda/2$
- The coax length must be short compared to the wavelength at the highest frequency

Equal delay transformers

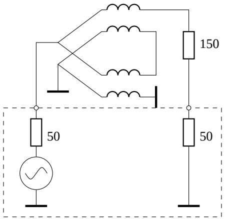
- These examples are also 1:4 transformers
- Signals travel the same distance, arrive in phase
- No more null in the response



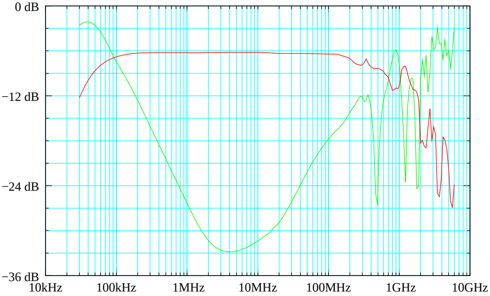


- *Very* wide bandwidths are possible
- Limited by leakage inductance and parasitic capacitance
- … and by residual length difference

Transmission line transformers

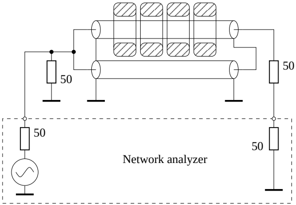


Test circuit for Guanella 1-4 transformer

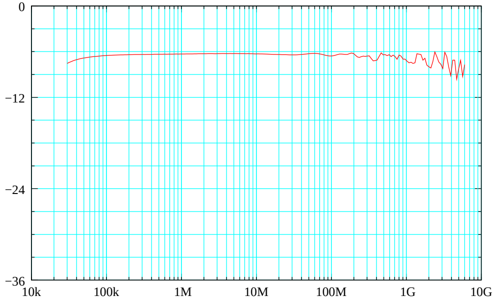


Frequency response of wire-wound Guanella 1-4 transformer

Transmission line transformers

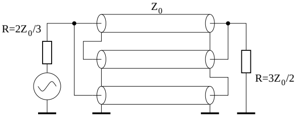


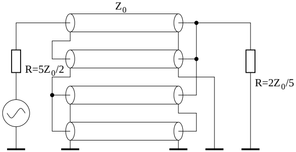
- Frequency response of a Guanella 1-4 transformer with coax



Equal delay transformers

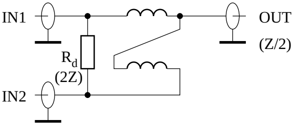
- What if you need ratios other than simple squared integers?

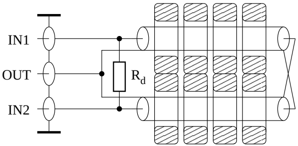




- Theoretically, all squares of rational numbers could be constructed
- In practice, the number of coax lines should remain small

Power combiners and splitters





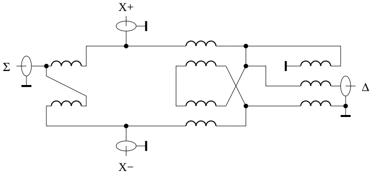
- This is an in-phase two-port combiner
- IN1 and IN2 are isolated from each other
- For good HF response, connections must be compact

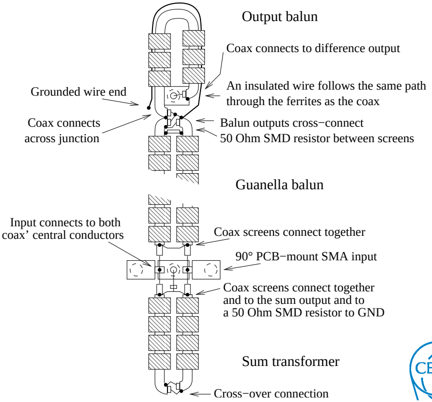
Power combiners and splitters

- A 180° two-port combiner (left) and a hybrid (right)
- IN1 and IN2 are isolated from each other
- For good HF response, connections must be compact

Hybrid transformers

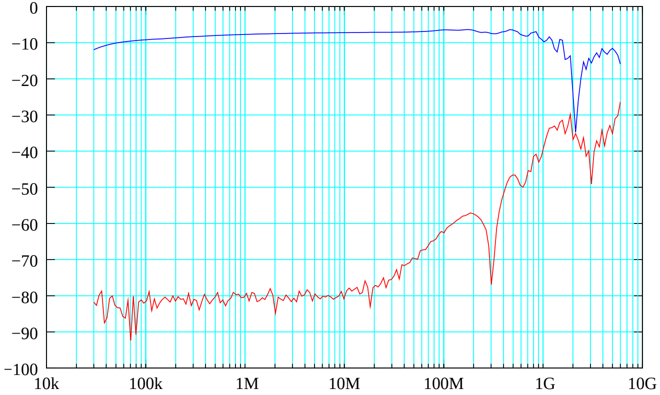
- Passive hybrid transformer for a 6 kHz-600 MHz beam position pick-up





Hybrid transformers

- Frequency response of Σ (top) and Δ (bottom) outputs with equal inputs



Hybrid transformers

- Photo of a 6 kHz-600 MHz hybrid transformer

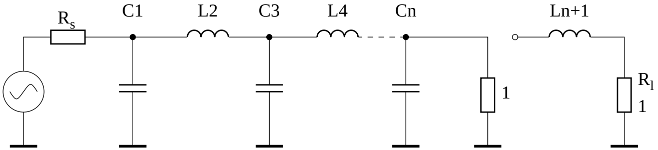
Passive LC filters

Passive LC filters

Why use passive LC filters?

- Reduce bandwidth
  - The interesting signal may span only a limited bandwidth
  - Restrict bandwidth prior to sampling, A-to-D conversion
  - Post-DAC reconstruction filter
- Reduce dynamic range
  - Some transducers deliver spikey signals, while all interesting information is in the baseband
- Reject out-of-band signals
  - Interference, other signal sources
- Reject out-of-band noise
  - Thermal noise

LC low-pass prototypes



- A sequence of LC sections
- May begin or end with either series L or parallel C
- The number of reactive elements is the *order* of the filter
- Stop-band energy is reflected
- Normalized load resistance: $R_l = 1$
- Normalized cut-off frequency $\Omega = 1$, (sometimes $F = 1$)
- … at half-power frequency (or sometimes at first ripple spec violation)

Filter families

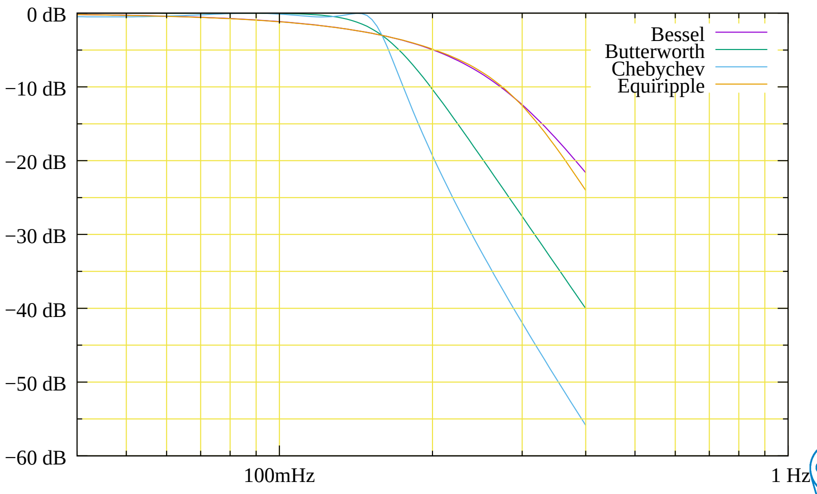
Optimized for:

- Flattest frequency response in pass-band (Butterworth)
- Linear phase response in pass-band (Bessel)
- Gaussian impulse response

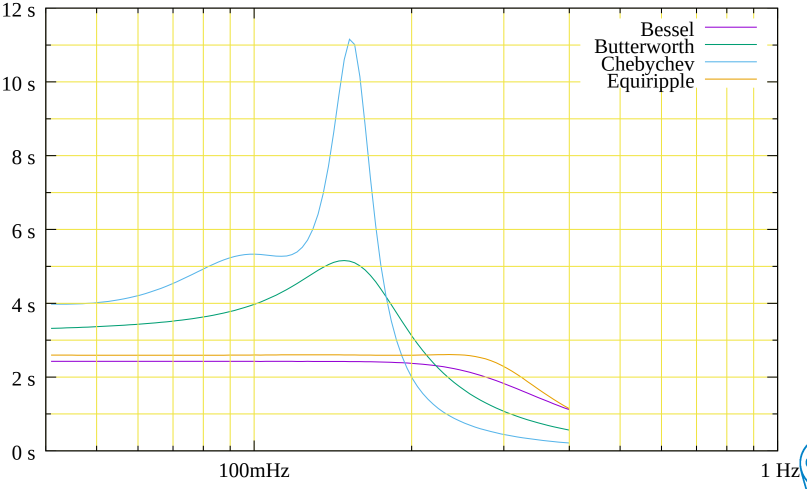
Compromise filters

- Brick-wall approximation, accepting some pass-band ripple (Chebyshev)
- Fastest transition from pass-band to stop-band, accepting some ripple and a limited stop-band attenuation (Elliptic or Cauer)
- Linear phase with equi-ripple
- … and other variations…

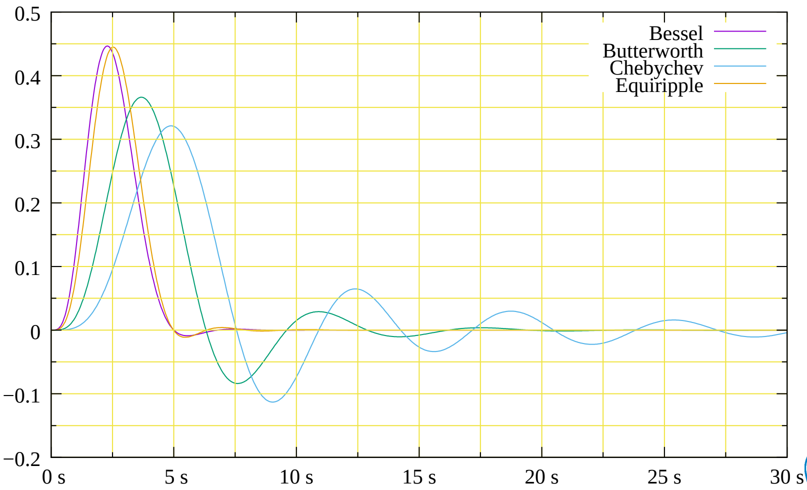
Frequency responses for some O(5) filters



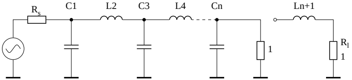
Group delay vs. Frequency for some O(5) filters

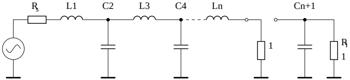


Impulse responses for some O(5) filters



Filter tables





Some normalized Bessel filter element values for $R_s = 1$

	C1 L1	L2 C2	C3 L3	L4 C4	C5 L5	L6 C6	C7 L7
2	0.5755	2.1478					
3	0.3374	0.9705	2.2034				
4	0.2334	0.6725	1.0815	2.2404			
5	0.1743	0.5072	0.8040	1.1110	2.2582		
6	0.1365	0.4002	0.6392	0.8538	1.1126	2.2645	
7	0.1106	0.3259	0.5249	0.7020	0.8690	1.1052	2.2659

Frequency and impedance scaling

The tabulated element values are basically the element impedances at the normalized load resistance and cut-off frequency.

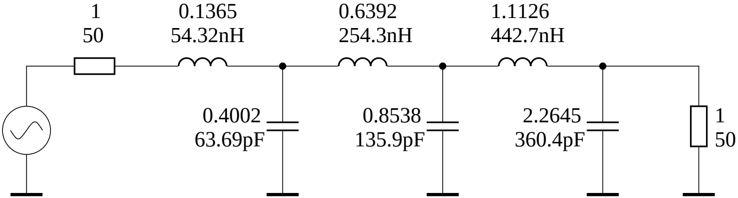
So the relations between the *real* and *normalized* values for target cut-off frequency ω and load impedance Z are:

$$C_r = \frac{C_n}{Z\omega} \qquad L_r = \frac{L_n Z}{\omega}$$

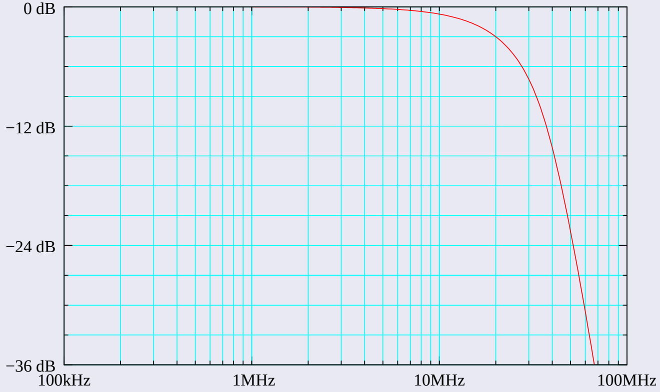
Example for a O(6) Bessel filter

Say: $Z = 50\ \Omega$ and $\omega = 2\pi * 20\ MHz$

- $C_r = 159.2p \cdot C_n$
- $L_r = 397.9n \cdot L_n$



Frequency response of the Bessel O(6) 20 MHz low-pass filter

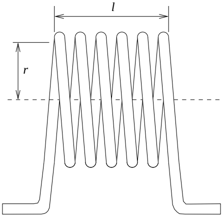


Finishing up the filter design

- You can't have 4-digit accurate inductors and capacitors.
- Common L's and C's have values in the E12 series ($\approx$ 20 % steps from one value to the next) and 5 % tolerances.
- You have to select from standard values.
- You may obtain a slightly better approximation by series or parallel combinations of two components but you'll still be limited by the basic component tolerances
- Depending on frequency and impedance choices, element values may end up impractically large or small

Making your own coils

- Don't shy away from making your own air core inductors!
- It's easy to get an accuracy much better than 5 %



$$L = \frac{\mu_0 \pi r^2 N^2}{0.9r + l}$$

- Aim for $l \approx 2r$
- Allow about one wire diameter of spacing between turns
- Good from $\approx$ 10 nH to 500 nH

An example LC filter realization

Bandpass filters

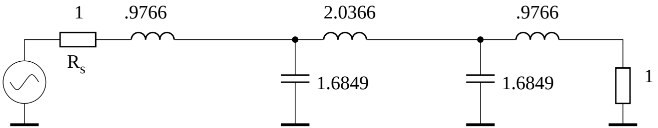
- The same filter element tables can be used to design bandpass filters
- You start off by designing a low-pass filter with a cut-off frequency at the target *bandwidth*.
- Then you replace each series component with a series L-C combination and each parallel component with a parallel L-C, both tuned to the desired centre frequency.

Example: A O(5) Chebyshev bandpass

Let's design an O(5) Chebyshev bandpass filter with 2 *MHz* bandwidth and 20 *MHz* centre frequency

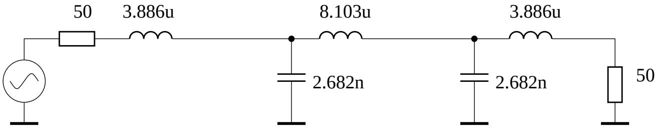
The normalized filter element values for $R_s = 1$

L1	C2	L3	C4	L5
0.9766	1.6849	2.0366	1.6849	0.9766

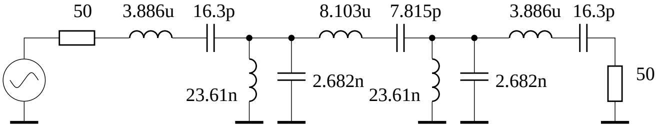


Example: An O(5) Chebyshev bandpass design example

Scale to 2 MHz and 50 Ω

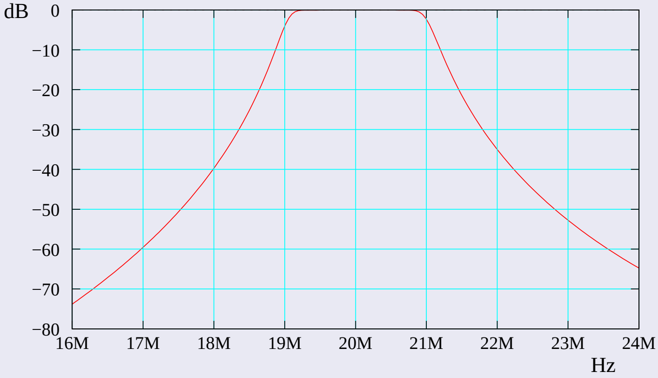


Resonate all elements to 20 MHz $\left(\frac{1}{\sqrt{LC}} = 2\pi \times 20\ MHz\right)$



Example: An O(5) Chebyshev bandpass design example

And the resulting frequency response plot:



Example: An O(5) Chebyshev bandpass design example

It's easy to end up with impractical element values

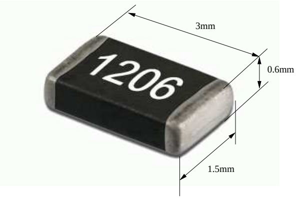
- It may be possible to arrange things using Norton's transform
- It may be possible to arrange things by applying star-delta transforms
- For very high frequencies, consider stripline filters
- For very low frequencies, consider active filters
- For very wide bandwidths, it may be easier to cascade a low-pass and a high-pass
- For very narrow bandwidths, there are other methods, involving weakly coupled staggered resonators, quartz, SAW, etc.

Intermezzo: Parasitics

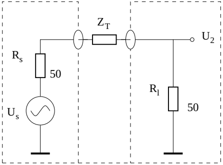
- Capacitance to floating nodes
- Capacitance and inductance of resistors
- Parasitic inductance and resistance of capacitors
- Self-capacitance and resistance in inductances
- Undesired inductive coupling

Resistors

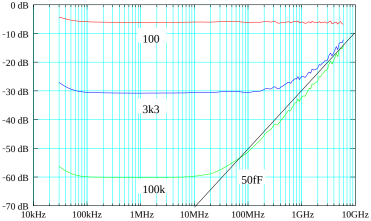
- Parasitics are rarely specified
- For SMDs, expect about 50 fF and 1 nH, almost independent of size and resistance
- MELFs often have a spiral cut $\rightarrow$ more inductance



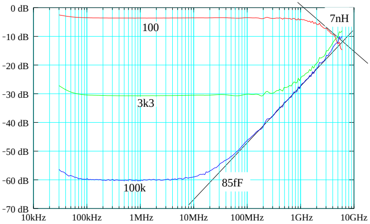
Resistor parasitics



- Setup to measure resistor parasitics



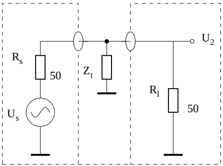
1206 SMD resistors



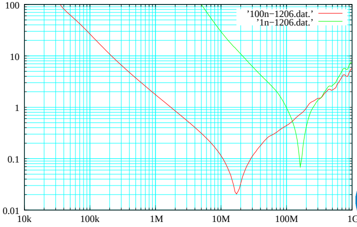
MiniMELF type resistors (1206 foot prints)

Capacitor parasitics

- 1206 SMD ceramic capacitors have about 1nH of inductance
- Very low losses and leakage for NP0 dielectric (small values)
- Large value capacitors use dielectrics that are non-linear, temperature-sensitive and hysteretic
- Some are even piezo-electric



Measurement setup



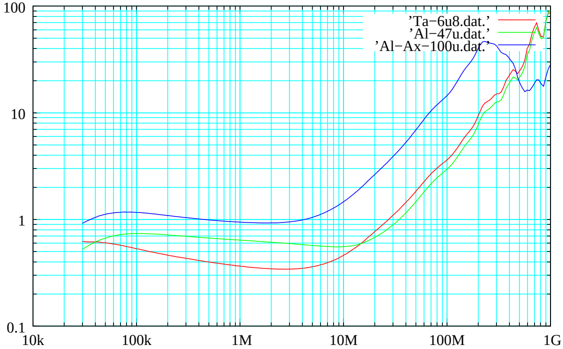
Impedance vs. frequency of some MLCCs

Electrolytic capacitor ESR ESL

Radial electrolytic: 5 nH, 500 mΩ

Axial electrolytic: 20 nH, 1 Ω

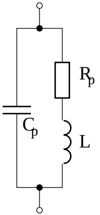
Ta electrolytic: 5 nH, 300 mΩ

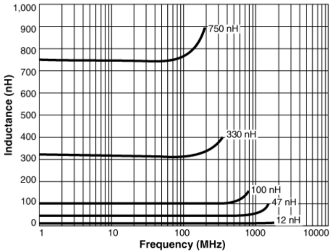


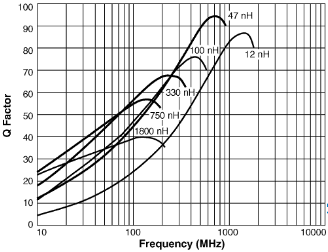
Impedance vs. frequency for some electrolytic capacitors

Inductor parasitics

- Wire resistance
- Distributed capacitance
- Skin effect: High-frequency current tends to flow in a thin surface layer
- External magnetic flux



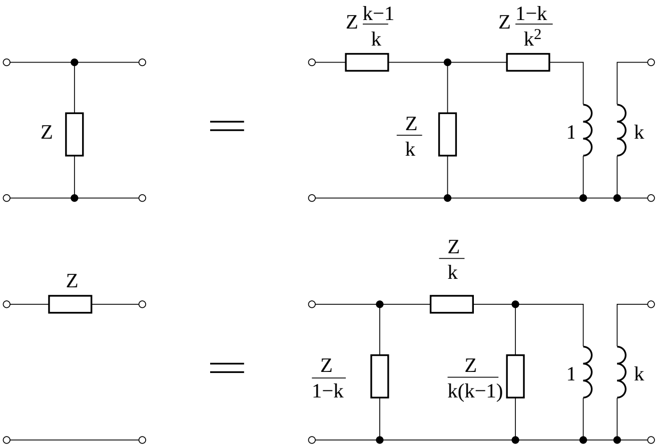




Plots from http://www.coilcraft.com

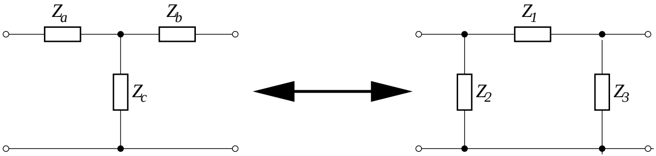
Back to passive Filters

Norton's transform



Note: k is the turns ratio of the ideal transformers

Star-Delta transform



$$Z_a = \frac{Z_1 Z_2}{Z_1 + Z_2 + Z_3}$$

$$Z_b = \frac{Z_1 Z_3}{Z_1 + Z_2 + Z_3}$$

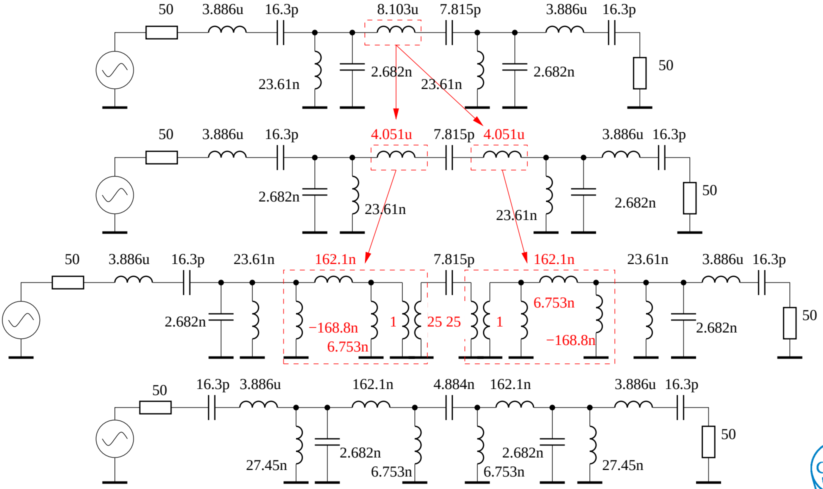
$$Z_c = \frac{Z_2 Z_3}{Z_1 + Z_2 + Z_3}$$

$$Z_1 = \frac{Z_a Z_b + Z_a Z_c + Z_b Z_c}{Z_c}$$

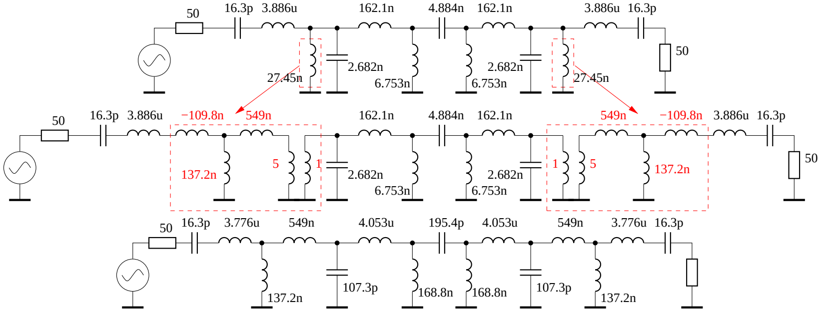
$$Z_2 = \frac{Z_a Z_b + Z_a Z_c + Z_b Z_c}{Z_b}$$

$$Z_3 = \frac{Z_a Z_b + Z_a Z_c + Z_b Z_c}{Z_a}$$

Applying Norton's transform to the O5 Chebychev BP filter



Applying Norton’s transform to the O5 Chebychev BP filter



Constant resistance filters

What's so special about Constant Resistance Filters?

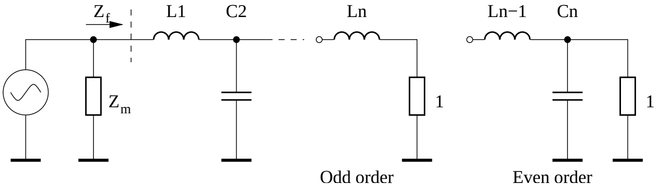
- They do not reflect
- They can be used to terminate long cables
- Frequency response does not depend on source resistance
- More complicated
- Only practical for some filter types:

- Butterworth
- Bessel
- Gaussian
- Almost, but not quite, for Linear Phase with Equiripple Error

Constant resistance filters

Principle

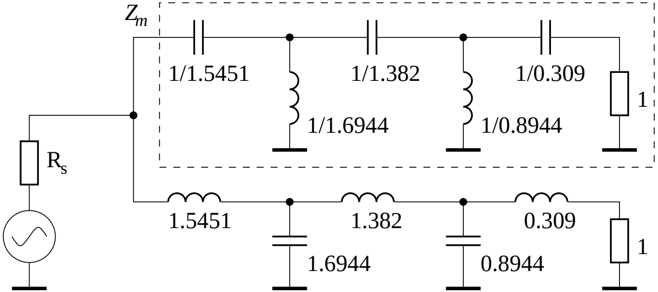
- Start with the normalized filter for zero source impedance
- Add a correcting (matching) impedance Z_m across the input



$Z_f \| Z_m = 1$

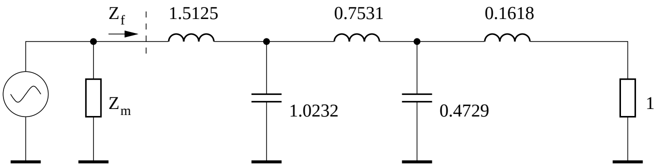
Constant resistance Butterworth filters

- The element values of Z_m are the *duals* of the main filter elements



Constant resistance Bessel filters

- The normalized filter element values for an O(5) Bessel for $R_s = 0$



$$Z_f = 1.5125s + \cfrac{1}{1.0232s + \cfrac{1}{0.7531s + \cfrac{1}{0.4729s + \cfrac{1}{0.1618s + 1}}}}$$

and

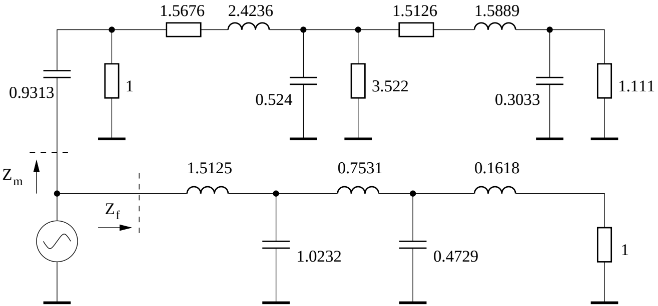
$$Y_m = \frac{1}{Z_m} = 1 - \frac{1}{Z_f}$$

Constant resistance Bessel filters

$$Y_m = \frac{0.9313s+1.60635s^2+1.22484s^3+0.4922s^4+0.0891777s^5}{1+2.4274s+2.61899s^2+1.58924s^3+0.55116s^4+0.0891777s^5}$$

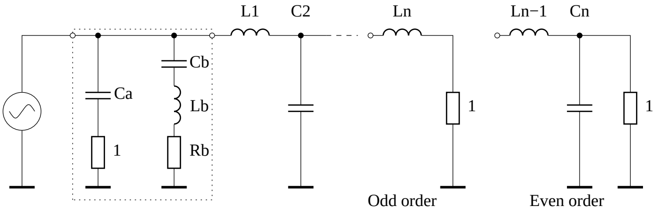
After continued-fraction expansion, we end up with:

$$Z_m = \frac{1}{0.9313s} + \cfrac{1}{1+\cfrac{1}{1.5676+2.4236s+\cfrac{1}{0.2839+0.524s+\cfrac{1}{1.5126+1.5889s+\cfrac{1}{0.8997+0.3033s}}}}}$$



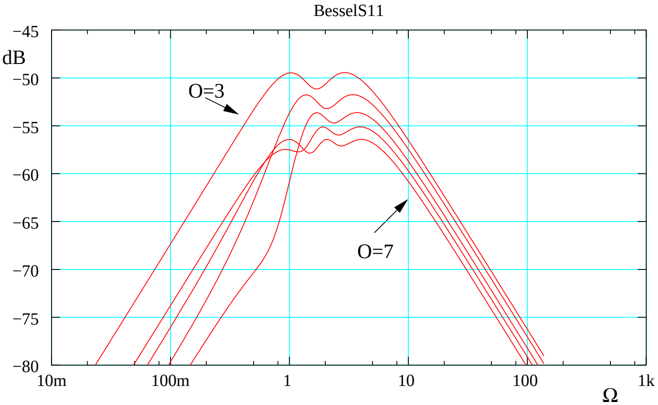
Constant resistance filters: Easier

- There is a simpler way
- The solution is not exact,
- ... but in practice it's plenty good



	Ca	Cb	Lb	Rb	L1	C2	L3	C4	L5	C6	L7
3	0.5804	0.3412	0.9915	2.6161	1.4631	0.8427	0.2926				
4	0.6121	0.3143	1.0646	2.7036	1.5012	0.9781	0.6127	0.2114			
5	0.6465	0.2834	1.1613	2.8896	1.5125	1.0232	0.7531	0.4729	0.1618		
6	0.6622	0.2683	1.2094	3.0029	1.5124	1.0329	0.8125	0.6072	0.3785	0.1287	
7	0.6876	0.2452	1.2955	3.2070	1.5087	1.0293	0.8345	0.6752	0.5031	0.3113	0.1054

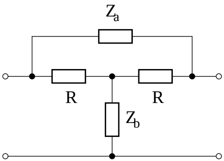
Reflection coefficient of some 'Easy' Constant Resistance Bessel LP filters

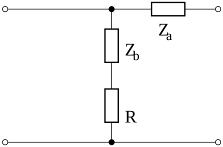


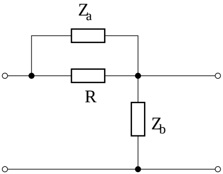
- It also works for Gaussian and equiripple phase error filters

Constant Resistance Networks

Constant resistance networks







- Z_a and Z_b are complex impedances such that $Z_a Z_b = R^2$
- The frequency response of the network is $\frac{R}{R+Z_a}$
- Load the right side with resistance R, and the left side will present a frequency-independent resistance R.

Constant resistance networks

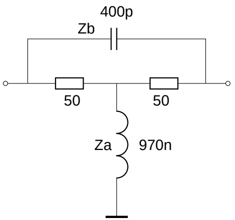
- Limited to one pole and/or one zero
- You can insert these networks in matched systems
- You can cascade these networks without interaction

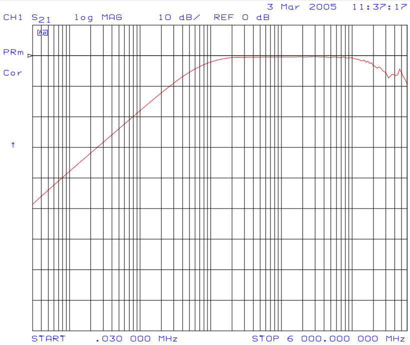
Applications:

- Frequency response correction (equalizers)
- Termination of out-of-band-signals
- Input impedance correction of amplifiers
- ...

Example

- A test jig for electrostatic PU amplifiers
- Simulates electrode frequency response





Noise in electronics

Noise

- By *noise* I mean undesired fluctuations intrinsic in a device
  - Thermal noise
  - Shot noise
- Undesired fluctuations coming from outside are *interference*
  - Radio frequency interference (RFI)
  - Power supply noise
  - ...

Thermal or Johnson noise

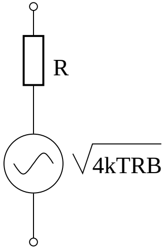
- Any device that converts electrical energy into heat also does the opposite
- In a bandwidth ΔB, a resistor delivers a noise power of: (Into a matched load)

$$P_n = kT\Delta B \text{ [W]}$$

- This noise is 'white' (Constant spectral density)
- This noise is Gaussian with $\mu_n = 0$
- It is as if the resistor had an internal voltage source:

$$e_n = \sqrt{4kTR} \ \ [\mathrm{V}/\sqrt{\mathrm{Hz}}]$$

$k = 13.8\mathrm{yW/HzK}$



Shot or Schottky noise

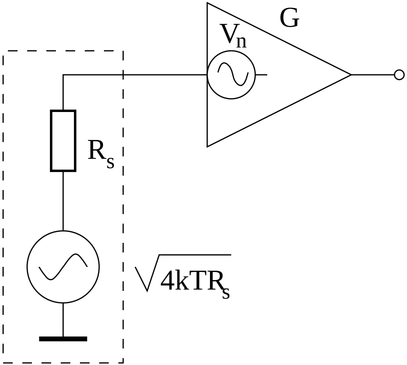
- Due to charge quantization
- Produced where a current flows across a potential barrier

$$I_n = \sqrt{2q_0 I_{dc}} \quad [\mathrm{A}/\sqrt{\mathrm{Hz}}]$$

- This noise is white
- This noise is Gaussian
- Metallic conductors have no Schottky noise

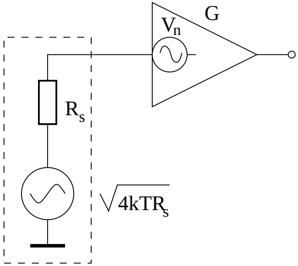
Noise in amplifiers

- It is customary to consider noise as if all of it originated at the amplifier input
- The term is "Input referred noise"
- That's actually close to being true, usually



Noise factor, noise figure

- The *noise factor* F is the ratio of total noise referred to the amplifier input, compared to the noise of the source alone
- Always greater than 1
- Usually reported in dB and then called 'Noise Figure': $NF = 10 \log F$

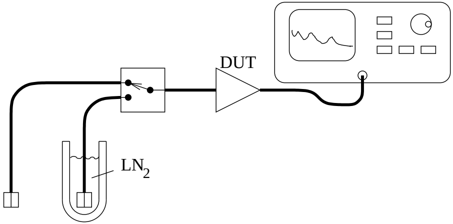


$$F = \frac{4kTR_s + {v_n}^2}{4kTR_s}$$

- Using this to get v_n is not very accurate

Measuring noise: The Y-method

- A noise generator with two well characterized output levels
- For example a 50 Ω terminator in LN_2 (77 K) and another at room temperature (296 K)
- We measure the amplifier's output noise change
- The amplifier's own noise tends to mask the change at the input.



Ratio of noise levels:
$10 \log \frac{296}{77} = 5.85$ dB

Measuring noise: The Y-method

- It's not easy to measure absolute noise levels
- ... but it *is* easy to measure a *change* in level
- We don't need an absolute calibration of the measurement instrument
- We don't need to know the gain of the amplifier
- The amplifier must have enough gain to overcome the measurement instrument's noise

Define Y as:

$$Y = \frac{P_a + P_h}{P_a + P_c}$$

Solve for P_a:

$$P_a = \frac{P_h - YP_c}{Y - 1}$$

Measuring noise: The Y-method

- From $P = U^2/R$, we can find V_n:

$$V_n = \sqrt{P_a R_{in}}$$

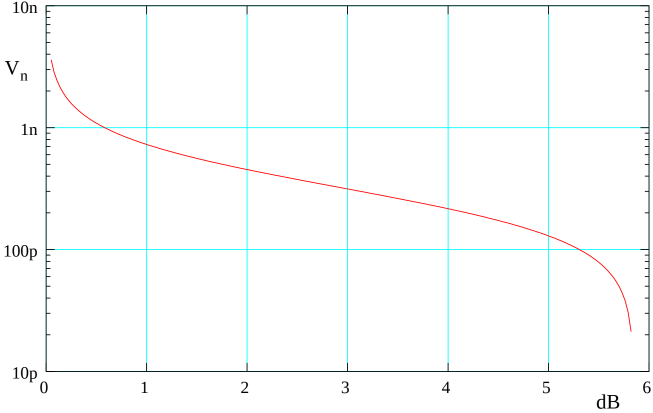
- and from $P = kT$ $(B = 1)$ we can derive an equivalent 'noise temperature':

$$T_n = \frac{P_a}{k}$$

- Note that attenuation in the path from the cold source *increases* its noise level
- This would make the amplifier look noisier than it really is

Measuring noise: The Y-method

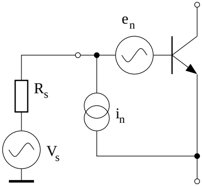
- For good accuracy, the noise generator's output should be in the same ballpark as the amplifier's own noise



V_n vs. Y

Noise in bipolar transistors

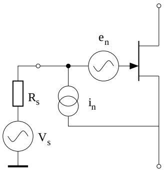
- Johnson noise from the base spreading resistance r_{bb}
- Collector current shot noise into the intrinsic emitter resistance $r_e = 1/g_m = kT/qI_c$
- Base current shot noise into r_{bb} (at low frequencies)



$$e_n = \sqrt{4kTr_{bb} + 2qI_c r_e^2 + \frac{2qI_c}{\beta} r_{bb}^2} \approx \sqrt{4kTr_{bb} + 2kTr_e}$$

Noise in FETs

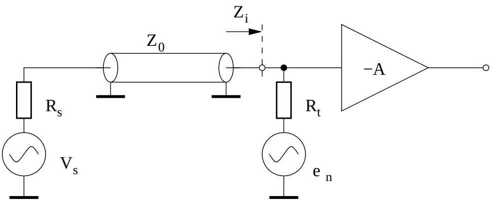
- Johnson noise of the channel resistance
- Schottky noise of the gate leakage current (Mostly irrelevant)



$$e_n = \sqrt{4kT\frac{2}{3g_m}}$$

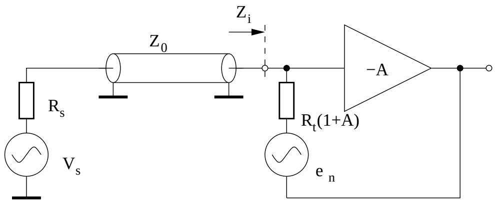
- For low e_n select JFETs with large g_m
- This implies large geometries and thus large capacitances

Impedance matching of LNAs



- Input referred noise voltage density due to R_t:
  $v_n = \sqrt{4kTR_t}\left(\frac{R_s}{R_s+R_t}\right) = \sqrt{kTR_t}$
- Not so great!

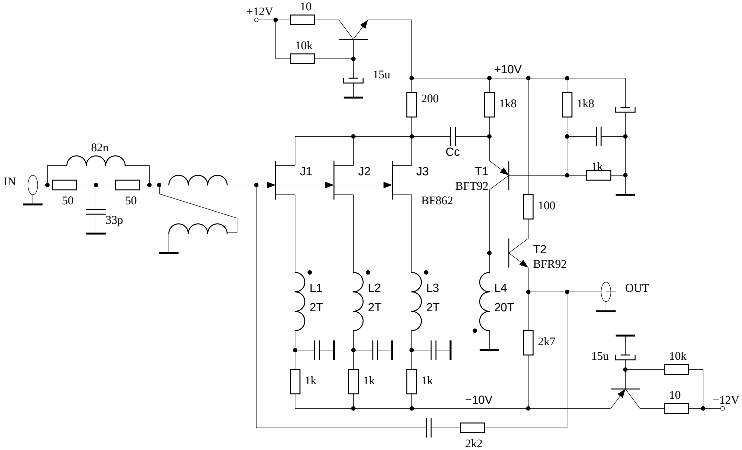
Impedance matching of LNAs



- Amplifier gain $-A$. Use largish A.
- To keep the same input impedance $R_t = (1+A)Z_0$
- Input referred noise voltage density $v_n = \sqrt{\frac{kTR_t}{1+A}}$
- Much lower noise!
- Phase shifts and gain errors in the amplifier will affect Z_i

A low noise amplifier design example

- A low-noise pre-amp design example
- G=26dB, $Z_i = 50\Omega$, $BW = 10\text{kHz-}30\text{MHz}$, $v_n = 260\text{pV}/\sqrt{\text{Hz}}$



Electromagnetic Interference

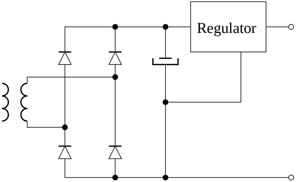
Electromagnetic interference

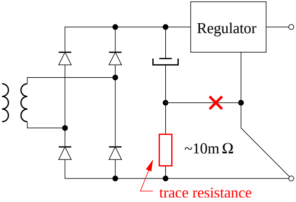
- Unwanted signals from outside leaking into your system
- Often difficult to fix:
  - The source is unknown
  - The coupling path is unknown
  - The critical components do not appear in any schematic diagram
  - ... and may not even *be* actual components

Coupling mechanisms

Common impedance coupling

- Do not share high current paths with low-level signals
- Use ground peninsulas or cuts (but don't get carried away)
- Star ground (LF only)

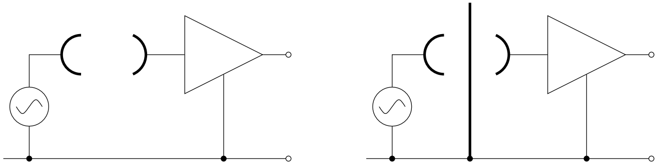




Coupling mechanisms

Electric field coupling:

- Affects high-impedance nodes
- Agressors are nodes with rapidly changing voltages with wide swings
- Use grounded or guarded shields
- Increase distance
- Lower victim node impedance

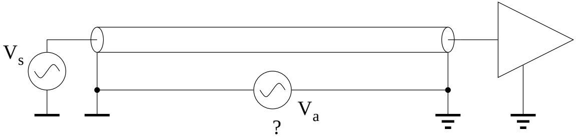


Magnetic coupling:

- Affects loops
- Keep loops with high currents small
- Keep victim loops small
- Put distance between them
- Screening is difficult

Coax cable leakage

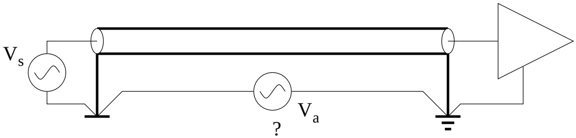
- A very common situation
- A coaxial cable connects two devices at different locations
- Some external agressor source imposes a potential difference
- Current flows in the coax screen
- Some of that leaks into the cable



- The screen's *purpose* is to conduct this current
- but some impedance is needed to limit it

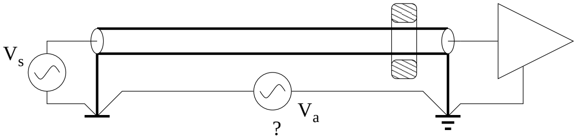
Coax cable leakage

- Install cable in grounded metal trays
- Use double-screened cable
- Pay attention to local grounding rules
- Never break the shield



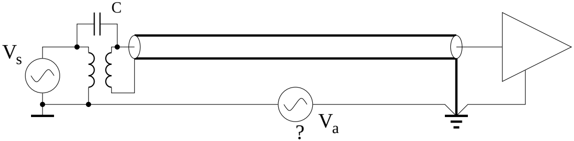
Coax cable leakage

- Increase common-mode inductance
  - Only useful for short connections
  - Not effective for low frequencies



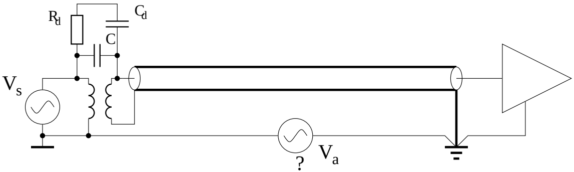
Coax cable leakage

- Separate grounds
  - Residual capacitance may resonate with common-mode inductance
  - Not effective at high frequencies



Coax cable leakage

- A damper network lowers the resonance frequency and damps the resonance
  - Choose $C_d > C_p$ and $R_d \approx Z_{C_p}$ at the resonance



Radiation effects

Radiation

How to choose materials

- Component survival
- Material activation
- Corrosive breakdown products
- Reliability level required
- Number of devices in use
- Ease of repair/access

Radiation

<10Gy/y

- Mostly safe

10Gy/y - 1k Gy/y

- Some electronics OK, maybe.
- Avoid PTFE or PVC insulation
- Avoid opto-couplers
- No lateral PNPs
- No local processors/controllers

Radiation

> 1kGy/y

- No PTFE! No PVC!
- No active electronics
- Ceramics and metals OK
- Glass fiber/epoxy components OK (e.g. FR4 PCBs)
- Ferrite and nanocrystalline magnetics OK
- Wire insulation PE, PEEK, Kapton OK.

Radiation tolerant electronic design

- The effects depend strongly on manufacturing details like geometry or doping profiles
- Even if some parameters go outside the specified range, this doesn't imply that a component is suddenly useless.
- 'Equivalent' devices of different makes may fare very differently. This may even happen for different lots of the same make!
- You can't know for sure if you haven't done the measurement

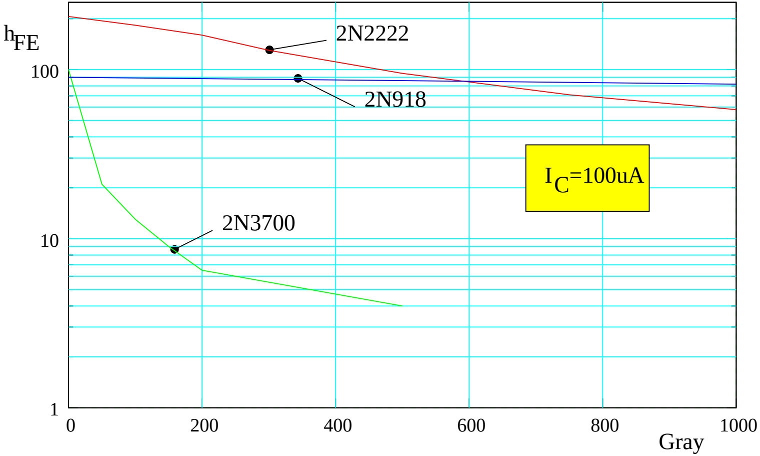
Radiation damage to...

Bipolar transistors

- Creation of recombination centers in the base
- Reduction of h_{FE} at low currents ($I_C < 100\mu$A)
- Design to tolerate wide variation in h_{FE}
- Use largish standing currents

Irradiated transistors lose current gain

Bipolar transistors will usually continue to work beyond 10kGy, but some do better than others.



Radiation damage to...

MOSFETs

- Ejection of e$^-$ from gate insulation layer
- V_{th} drifts downward
- Design to tolerate large variation in V_{th}

JFETs

- Increased gate leakage
- Increased noise, especially below 100 kHz

Use feedback to stabilize working points

Radiation damage to...

Linear integrated circuits

- NPN-only circuits are mostly robust (>1 kGy)
- Lateral and substrate PNP transistors are very susceptible (<100 Gy)
- Amplifier and comparator input bias currents tend to rise

- LM317 survives several kGy, but LM337 dies <100 Gy
- LF351 OpAmps still work with more than 10 kGy accumulated dose.

Radiation damage to...

Logic

- ECL and old TTL are quite radiation resistant ($>$ 1kGy)
- More recent logic is much more susceptible ($<$ 30Gy sometimes!)
- Use only simple logic, state machines and registers
- Beware of Single Event Upsets:
- Rewrite data frequently from a remote location
- Design state machines free of lock-up states
- Use redundant circuitry

Old-fashioned TTL, 74S, 74LS seem to hold up well beyond 1 kGy, but 74F dies at less than 100 Gy. EPM7064 (EEPROM) FPGAs seem to survive well, but I have none that were exposed to more than an estimated 500 Gy.

Something will break/drift/change

Whether a device is operational or not depends much on how its components are used. If correct operation relies on a parameter that happens to drift under irradiation, your circuit dies early.

Allow for parameter drift

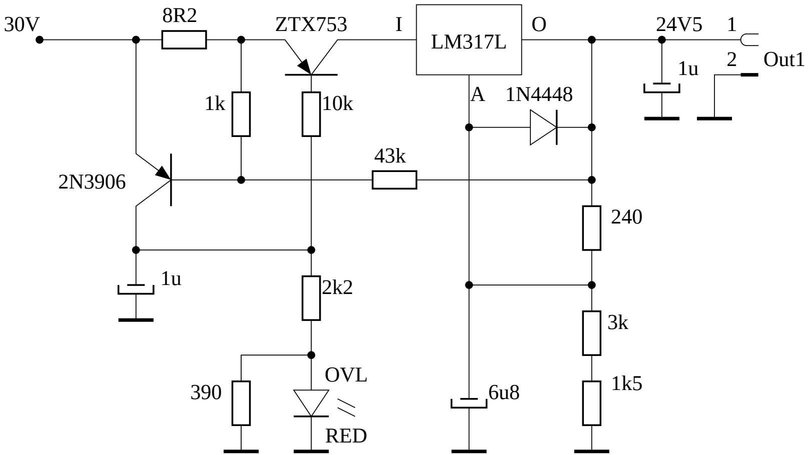
- Allow for large changes in bias/leakage currents, V_T, h_{FE}
- Avoid very high impedances
- Use largish standing currents
- Avoid ICs containing lateral or substrate PNP transistors

Defensive Design

Try to confine damage

- Remote power supplies (Easy to clear latch-ups, too!)
- Split power distribution
- Fold-back current limiting, PTC or PolyFuse
- Insert sense resistors in power supply connections

Example: A power supply with fold-back current limiting



Thank you