Basics of RF Electronics

Lecture 1

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Copyright statement and speaker's release for video publishing

Plan

Lecture 1

- Components
- Transmission line theory re-cap
- Connectors
- Reflection
- Co-planar waveguide
- Surface mount components
- Printed structures
- Filters
- Kuroda identities
- Wilkinson splitter

Lecture 2

- Mixers
- IQ modulation
- Oscillators
- Phase noise
- Phase locked loops
- Learning resources

RF Electronic Components

Specialised RF components and devices include

- transmission lines
- connectors
- IQ modulators
- phase detectors
- amplitude detectors
- oscillators
- RF switches
- frequency dividers
- frequency multipliers
- voltage controller oscillators
- amplifiers
- mixers
- antennae
- filters
- couplers
- circulators
- Splitters
- ADC & DAC
- FPGA

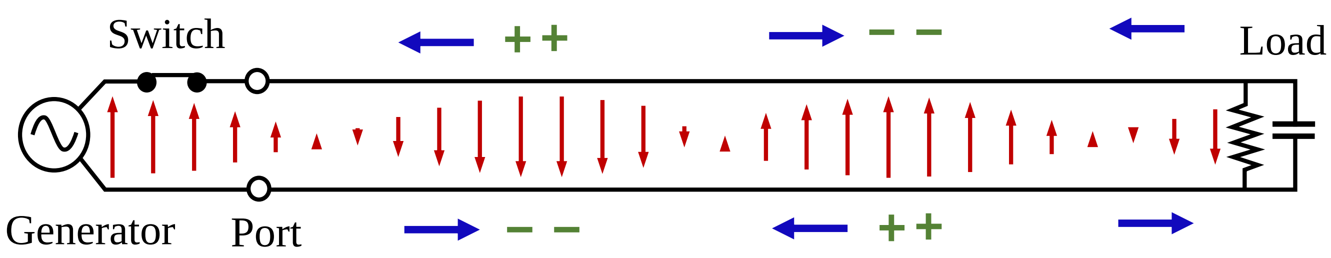
Passive components can be printed directly onto a PCB. Active components or precision passive devices might be supplied in metal cases with coaxial connectors or as surface mount devices to go on a PCB

Suppliers

Mini-circuits, Analog Devices, Texas instruments, MITEQ, Fairview, Pasternack, Microsemi, Infineon, …..

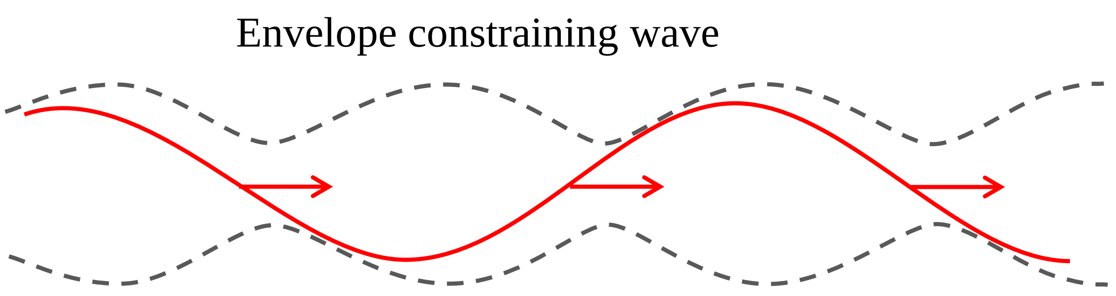
Transmission Lines

A distinction between AC theory and RF electronics is the need to consider many circuit interconnects as transmission lines.



Coaxial cable and microstrip are examples of transmission lines

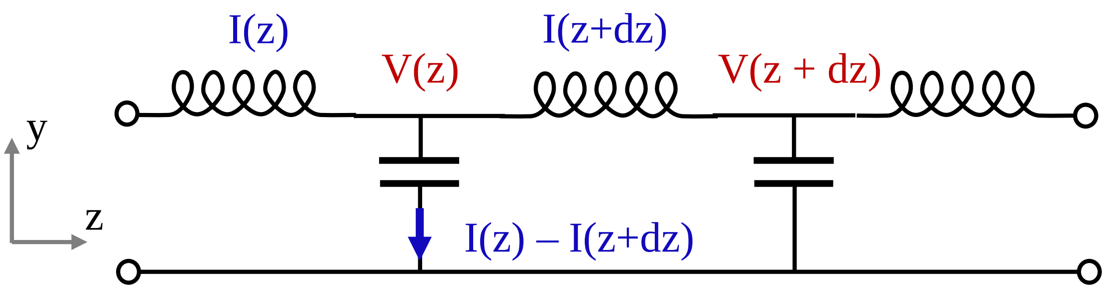
After the switch is closed, the EM field that starts to propagate has no knowledge of the line termination. The wave front of a TEM mode propagates at the speed of light for the material around the conductors.



Reflection from the termination gives a wave that surges through an envelope defining the standing wave ratio (VSWR).

For a forward wave (TEM mode) the current is in phase with the electric field. At full reflection VSWR $\rightarrow \infty$ and the current is 90^o out phase with the electric field.

Transmission Line Theory



Electronic engineers like to solve RF problems with circuits rather than solving Maxwell's equations

Circuit equations $\frac{\partial V}{\partial z} = -L\frac{\partial I}{\partial t}$ $\frac{\partial I}{\partial z} = -C\frac{\partial V}{\partial t}$ gives $\frac{\partial^2 V}{\partial z^2} = LC\frac{\partial^2 V}{\partial t^2}$ set $c = \frac{1}{\sqrt{LC}}$

Where C is capacitance per unit length, L is inductance per unit length, c is the phase velocity

Solutions of the wave equation can be decomposed into forward and backwards waves as

$$V(z,t) = F\left(\frac{t}{\sqrt{LC}} - z\right) + R\left(\frac{t}{\sqrt{LC}} + z\right)$$

Where F is information travelling to the right and R is information travelling to the left

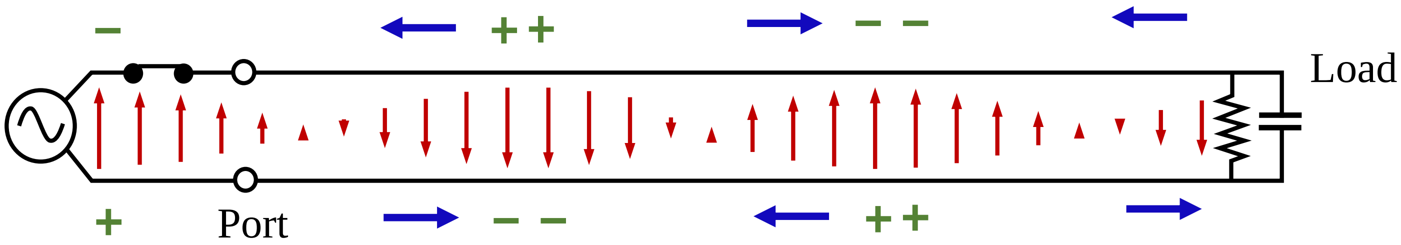
Using either one of the circuit equation above gives the current as

$$I(z,t) = \sqrt{\frac{C}{L}}\left\{F\left(\frac{t}{\sqrt{LC}} - z\right) - R\left(\frac{t}{\sqrt{LC}} + z\right)\right\}$$

The ratio of voltage to current for a forward wave without reflection gives the intrinsic impedance as

$$Z_o = \sqrt{\frac{L}{C}}$$

Ports



The figure marks a port. When considering RF circuits, it is convenient to consider sections independently. As the phase on the earth depends on location, one cannot complete circuit diagrams with arbitrary connections to earth.

Sections of RF circuit are defined by pairs of terminals forming a port. The condition for a pair of terminals to constitute a port is that the current entering one equals the current leaving the other in both phase and amplitude.

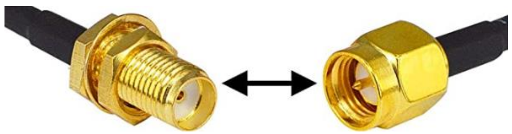
For connections between LLRF circuits on accelerators one invariably choses co-axial cable with intrinsic impedance of 50 Ω.

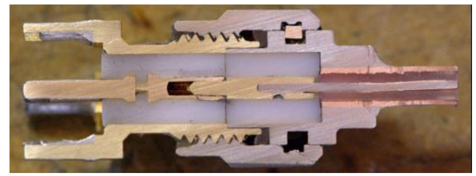
This value was adopted as a standard for instrumentation, being a compromise between minimizing attenuation and maximizing power transfer [2].

The power handling can be increased by increasing the diameters of the conductors subject to one ability to cool the inner conductor. The useful limit for measurement systems is the dimension where over-moding occurs at the frequency of interest.

Transmission Line Reflection

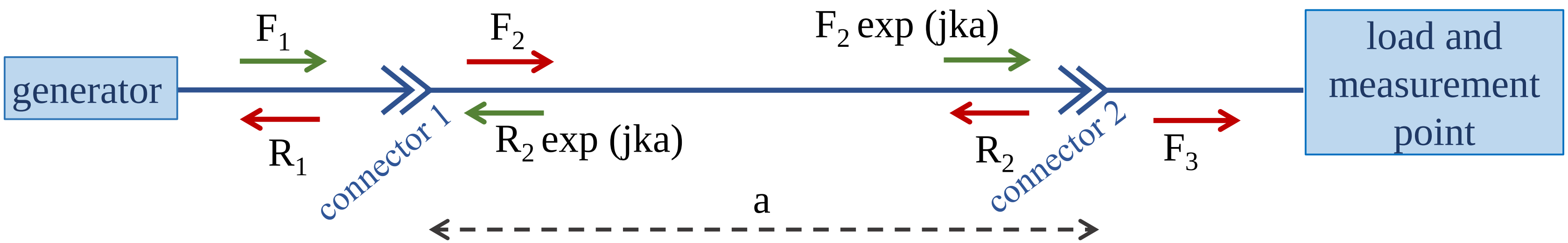
Measuring and preserving the phase of a signal can be critical to the ultimate performance of an accelerator. Unmatched components, changes in the transmission line type (e.g. microstrip to coax) and imperfect connectors give unwanted reflection.





[3]

Simple analysis can be undertaken for two connectors



Taking S matrices as unitary, and dependent on a single real variable δ providing a slight perturbation from the anti-identity matrix, then the system equations are

$$\begin{bmatrix} R_1 \\ F_2 \end{bmatrix} = \begin{bmatrix} j\delta & \sqrt{1-\delta^2} \\ \sqrt{1-\delta^2} & j\delta \end{bmatrix} \begin{bmatrix} F_1 \\ R_2 \exp(jka) \end{bmatrix} \qquad \begin{bmatrix} R_2 \\ F_3 \end{bmatrix} = \begin{bmatrix} j\delta & \sqrt{1-\delta^2} \\ \sqrt{1-\delta^2} & j\delta \end{bmatrix} \begin{bmatrix} F_2 \exp(jka) \\ 0 \end{bmatrix}$$

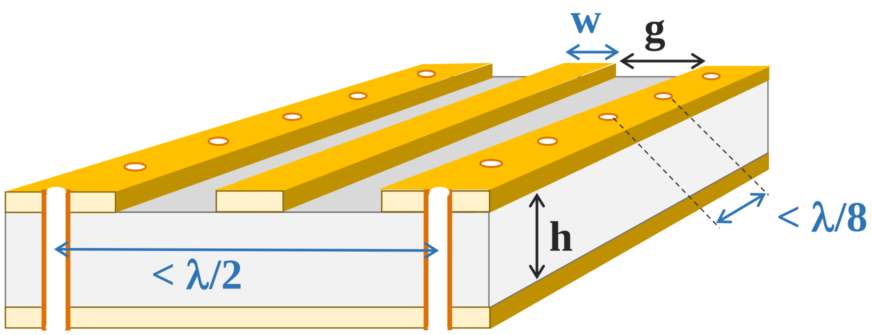
Solving gives $F_3 = \dfrac{1-\delta^2}{1+\delta^2 \exp(2jka)} F_1 \exp(jka)$ Zero phase shift when $2ka = n\pi$

Co-planar waveguide

PCBs have printed copper tracks on a substrate. Multilayer boards with internal tracks are formed by gluing two-layer boards together.

At high frequencies the tracks must be designed as transmission lines.

A type of transmission line that is widely used with PCBs is grounded coplanar waveguide (GCPW)



Closely space vias (conducting paths) connect the upper and lower earth planes.

Varying gap between the strip and adjacent earth planes adjusts impedance over a small range independently of strip width allowing strip width to be adjusted to that of surface mount components.

A good approximation for the impedance is obtained using the following equations

$$k_1 = w + 2g \qquad k_2 = \sqrt{1 - k_1^2} \qquad k_3 = \frac{\tanh\{0.25\pi w/h\}}{\tanh\{0.25\pi(w+2g)/h\}} \qquad k_4 = \sqrt{1 - k_3^2}$$

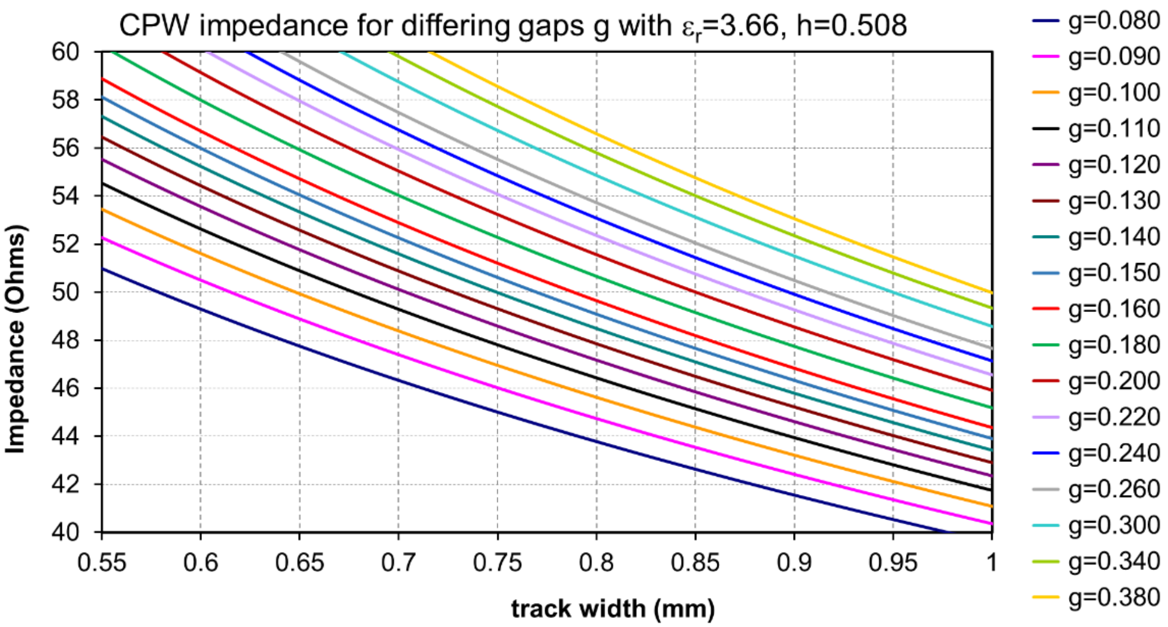
$$\varepsilon_{eff} = \frac{K(k_1)K(k_4) + \varepsilon_r K(k_2)K(k_3)}{K(k_1)K(k_4) + K(k_2)K(k_3)} \qquad Z_o = \frac{60\pi}{\sqrt{\varepsilon_{eff}}} \frac{K(k_2)K(k_4)}{K(k_1)K(k_4) + K(k_2)K(k_3)}$$

$K(k)$ is the complete elliptic integral of the first kind

Ref [11]

CPW Impedance

Using the formula, GCPW track impedance of substrate RO4350B with thickness 0.506 mm is determined as a function of track width and for differing gaps. For example, the graph gives a 50 Ω impedance for track width 0.85 mm with a gap of 0.2 mm. The formula does not account for track thickness and indeed more accurate results would be given by direct simulation.

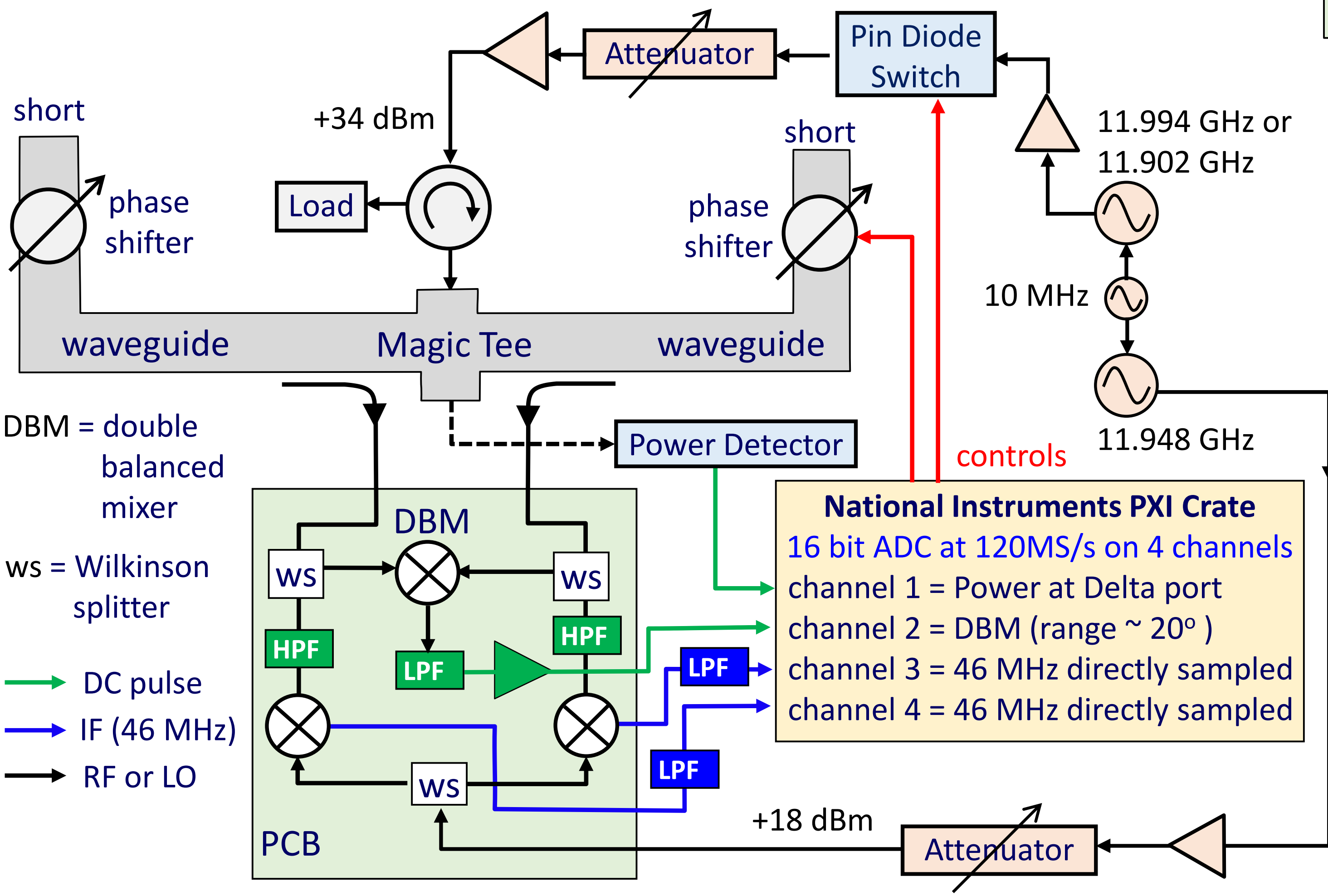


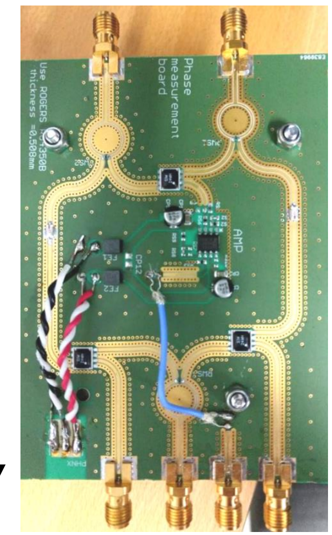
Up to a few GHz the substrate FR-4 is commonly used. It is flame retardant and made from woven glass fibre, epoxy resin and a mineral filler. Its high frequency properties are not tightly controlled or specified, above 6 GHz losses are usually unacceptable. Special low loss substates such as the Rogers RO4000 series should be used at high frequencies. For RO4350B the manufacturers quote a loss factor of 0.0037 and a design relative permittivity (Dk) of 3.66 at 10 GHz. [4]

Test PCB for a synchronisation experiment

CLIC has a requirement to synchronise the phase two crab cavities to ~ 5 fs [10]. RF pulse is 156 ns so no time for feedback. Effective waveguide lengths must be controlled to a 1 micron in length over two 35 metre runs.

A PCB was fabricated to test the concept.





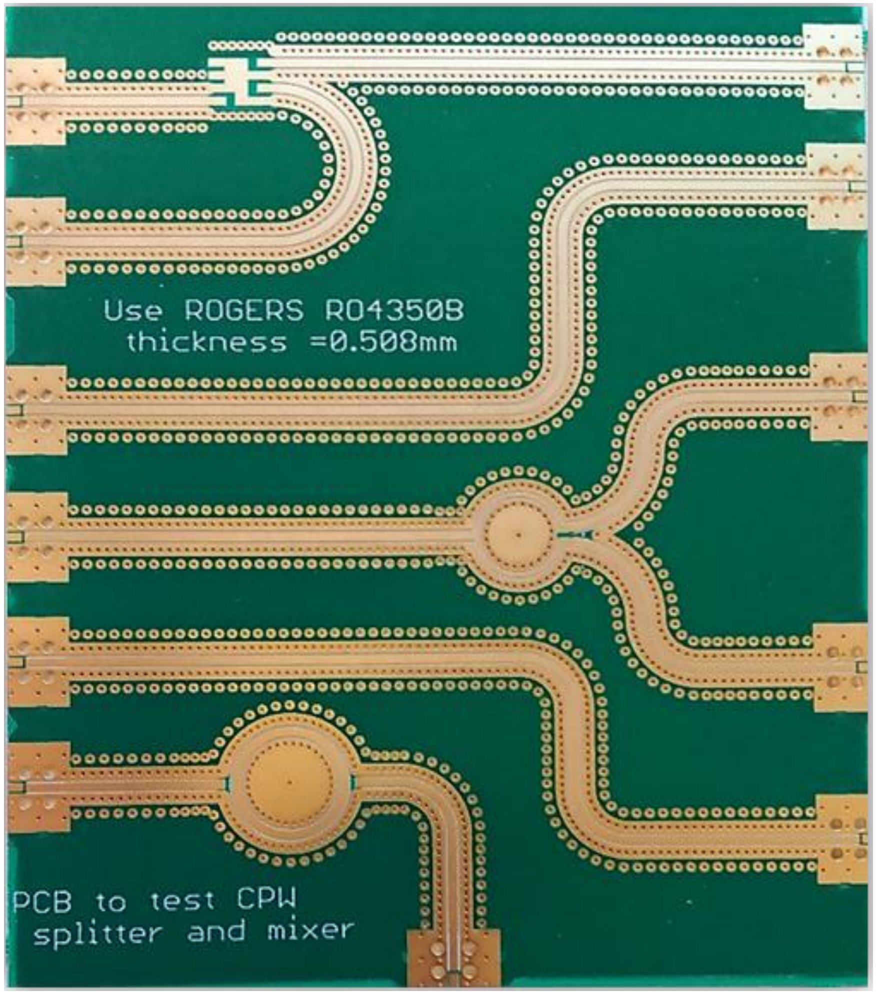
Prior a test PCB was fabricated to develop the CPWG

GCPW test board

Mixer test

Two rows of vias are used to bond top earth to ground plane

Resonator to check dielectric constant



Single path with two bends

Test for Wilkinson splitter

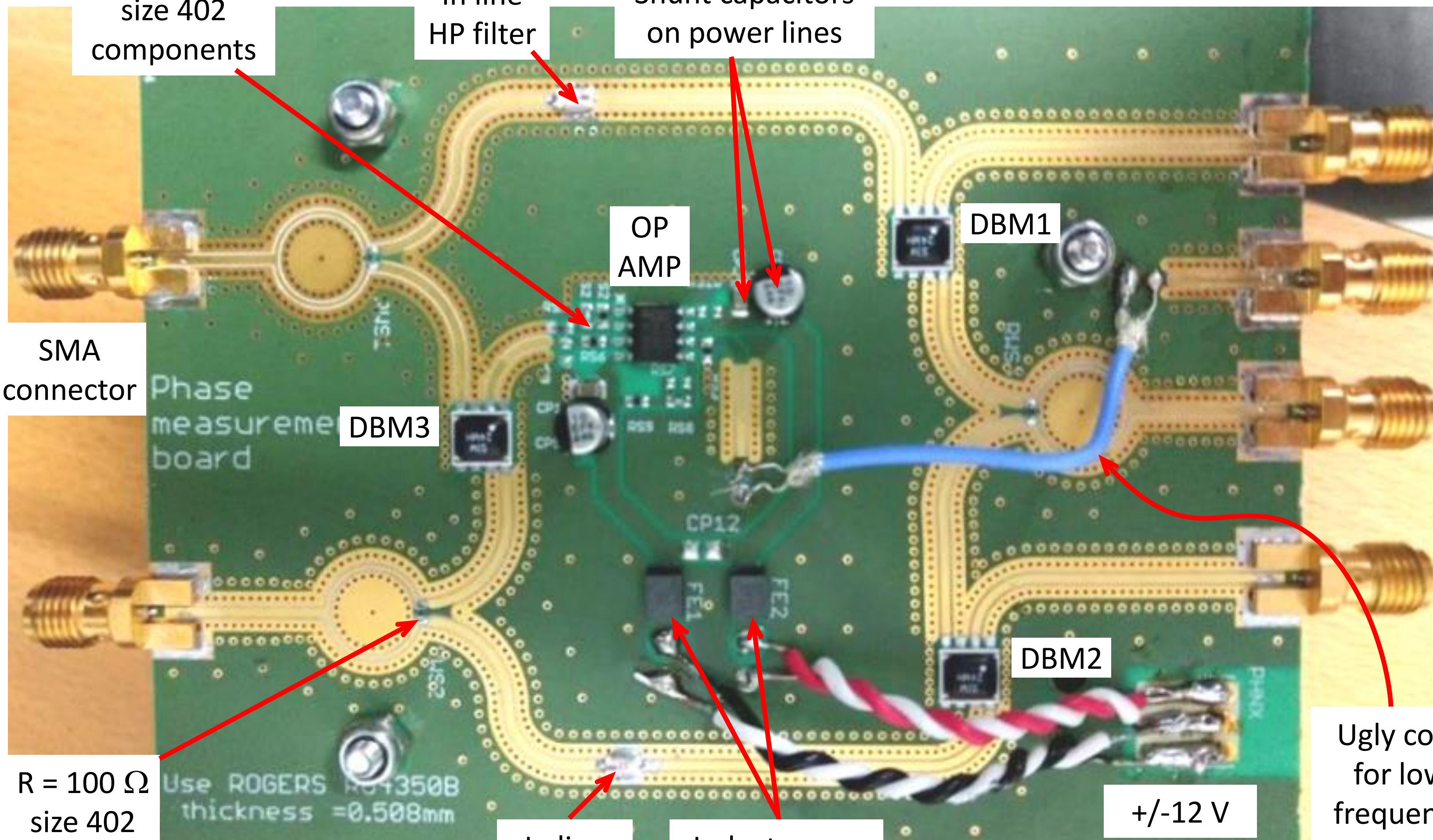
Green layer is the solder resist. It was not coated on the tracks, the gap or the edge of the top earth

Single path with two bends minus one quarter of a wavelength compared to above

After analysis of transmission for tracks differing by ¼ wavelength, the required design permittivity for this fabrication using the formula above was close to 3.545 rather than 3.66. With this revised permittivity the required track width for a gap of 0.2 mm became 0.789 mm.

Examples of surface mount components

The CERN Accelerator School



DBM = Double Balanced Mixer
DBM1 and DBM2 down-convert to the intermediate frequency (IF)

Two layer board – flying leads are avoided with multilayer boards

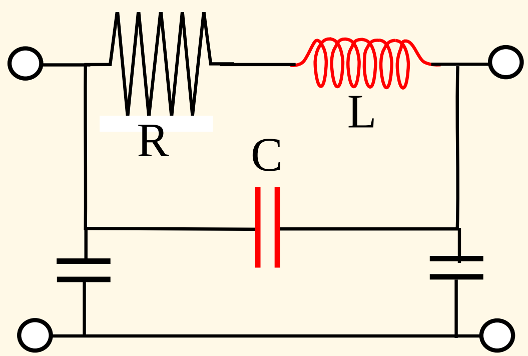
Surface mount components

Case length (1/1000 inch)	Case length (mm)	Case width (mm)	Gap between end caps (mm)
0201	0.6	0.3	0.16
0402	1.005	0.5	0.5
0603	1.608	0.8	1.0
0805	2.012	1.25	1.3
1206	3.20	1.60	2.4

Size specifications may be metric or imperial (imperial here)
Try to match component width to track width for series components.
Size 0201 needs to be placed under a microscope with a manipulator. Bigger components handle more power.

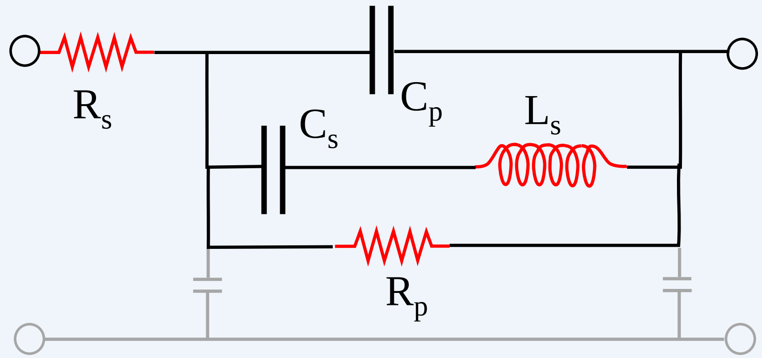
High Frequency model for a resistor

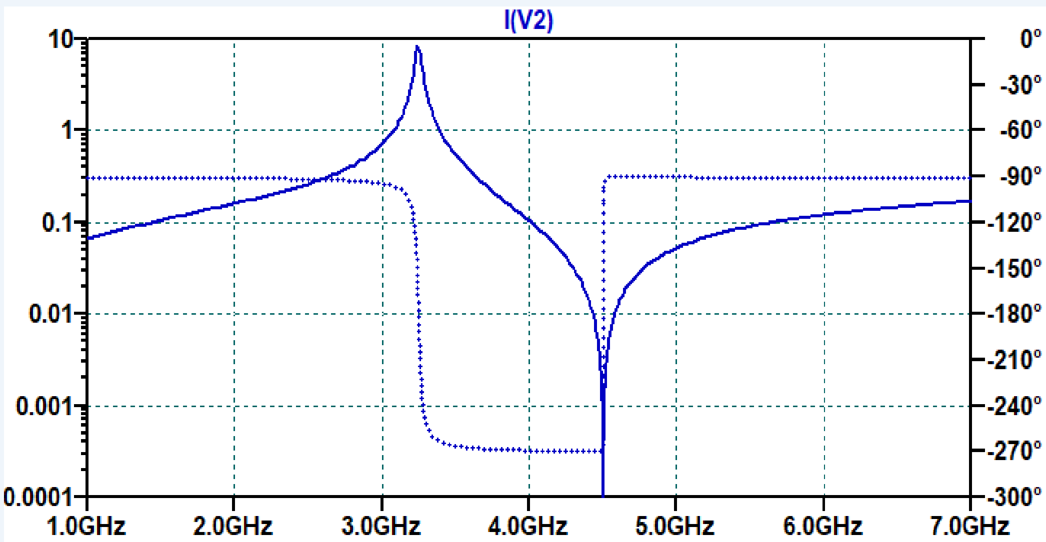
At high frequencies components have unwanted parasitics associated with leads, ground planes, lossy dielectrics and geometry.



Vishay thick film commodity resistor [6]
Failure rate < 10^{-9} h^{-1}

Capacitance model





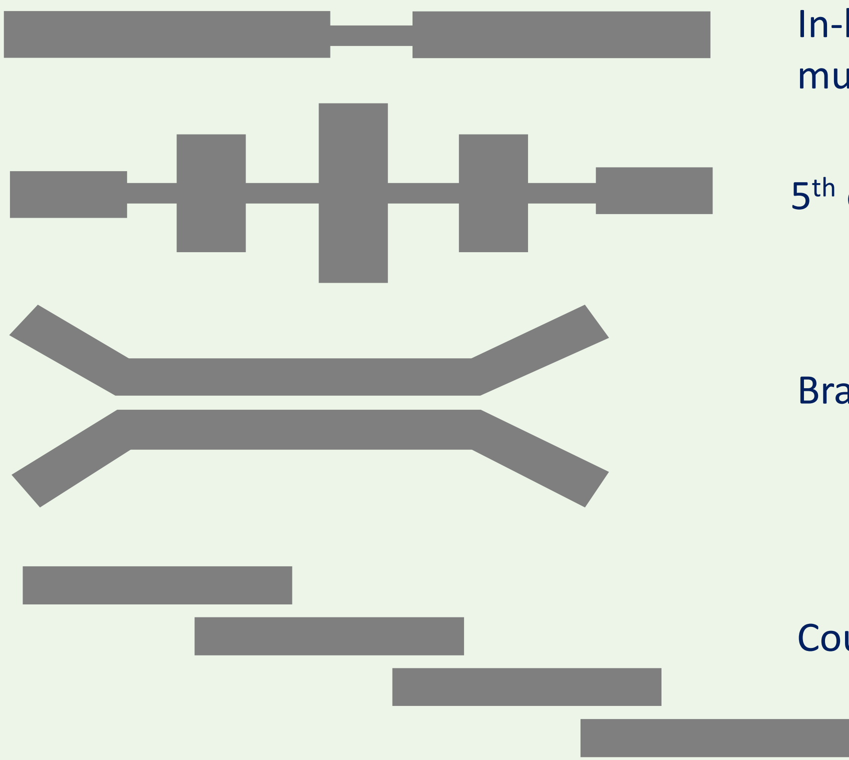
An example [5] from Johanson Technology for a 10 pF capacitor at low frequencies, is Cs = 4.8 pF, Cp = 5.2 pF, Ls =0.5 nH, Rs = 0.11 Ω and Rp = 78 kΩ. The current for an applied voltage of one is plotted Below 1 GHz the current rises as jωC. Above 1 GHz the effective capacitance starts to rise above 10 pF. The series resonance is at 3.25 GHz and at this point the current would be 9 A (if 1 Volt could be maintained)

Conductors forming fine layers of capacitors have inductance hence some part of the overall capacitance is in series with this inductance. The equivalent circuit drawn has a series LCR circuit with C_s and L_s and a parallel LCR circuit with C_p C_s L_s and R_p

Printed Structures

Passive structures can be printed on to a PCB to act as inductors, small value capacitors and to undertake a few important functions, including:-

- Stubs for impedance matching
- Filters (Low pass, High Pass, Band Pass, Band-stop)
- Splitters (Wilkinson)
- Couplers (Directional, Branch line, Hybrid Ring)
- Resonators



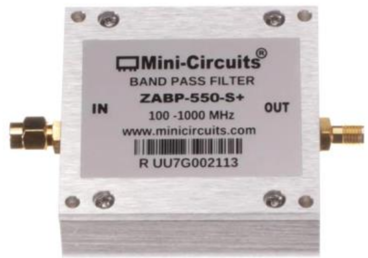
In-line inductor (OK for small values, must be much shorter than shortest wavelength.)

5th order low pass filter

Branch line directional coupler

Coupled resonator band pass filter

Connectorized Components



Designing a structure giving an S-Matrix to a high degree of accuracy requires considerable effort.

Accelerator system engineers often choose connectorized components.

Dissecting passive connectorized components often reveals suspended strip-line where microstrip tracks are mirrored and there is dielectric and an earth plane above and below the pair of mirrored tracks making the performance similar to a standard strip-line.

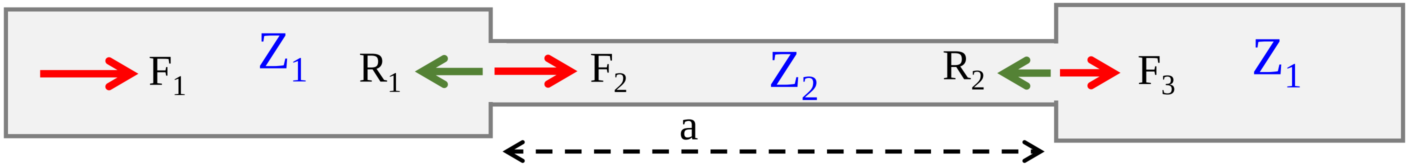
Advantages of strip-line are lower losses hence higher Q structures and reduced cross talk between tracks.

Design equations for fabricating passive structures exist for all the preferred transmission lines including microstrip, co-planar wave guide and strip-line.

Precise design invariably needs EM simulation and testing for the substrate and transmission line of interest, bespoke EM design tools include Keysight Pathwave Advanced Design System (ADS), Cadence AWR Microwave Studio, Altium Designer, Dassault CST Studio Suite, Siemens (PADS), ANSYS HFSS and Ansoft Designer

Impedance discontinuity

Representing a connector or transmission line imperfection.



At a change of transmission line impedance, voltages for incoming and outgoing waves balance.

Current is conserved across the junction.

From slide (6) $\quad F_1 + R_1 = F_2 \qquad (F_1 - R_1)/Z_1 = F_2/Z_2$

Solving gives $\quad R_1 = \left(\dfrac{Z_2 - Z_1}{Z_1 + Z_2}\right) F_1 \quad$ (18.1) $\qquad F_2 = \left(\dfrac{2Z_2}{Z_1 + Z_2}\right) F_1 \quad$ (18.2)

Considering both discontinuities in diagram. Assume outgoing transmission line has a reflectionless termination and sinusoidal wave with wavenumber k.

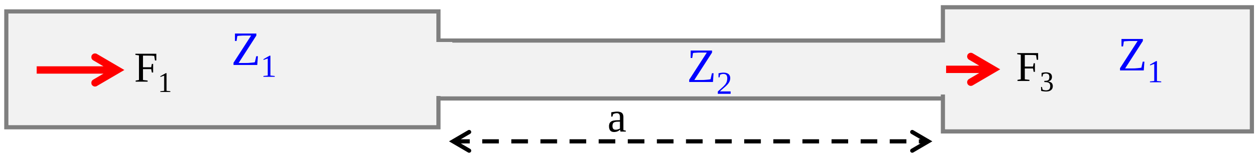
$$R_2 = \left(\frac{Z_1 - Z_2}{Z_1 + Z_2}\right) F_2\, e^{jka} \qquad F_2 = \left(\frac{2Z_2}{Z_1 + Z_2}\right) F_1 + \left(\frac{Z_1 - Z_2}{Z_1 + Z_2}\right) R_2 e^{jka}$$

$$F_3 = \left(\frac{2Z_1}{Z_1 + Z_2}\right) F_2\, e^{jka} \qquad R_1 = \left(\frac{Z_2 - Z_1}{Z_1 + Z_2}\right) F_1 + \left(\frac{2Z_1}{Z_1 + Z_2}\right) R_2\, e^{jka}$$

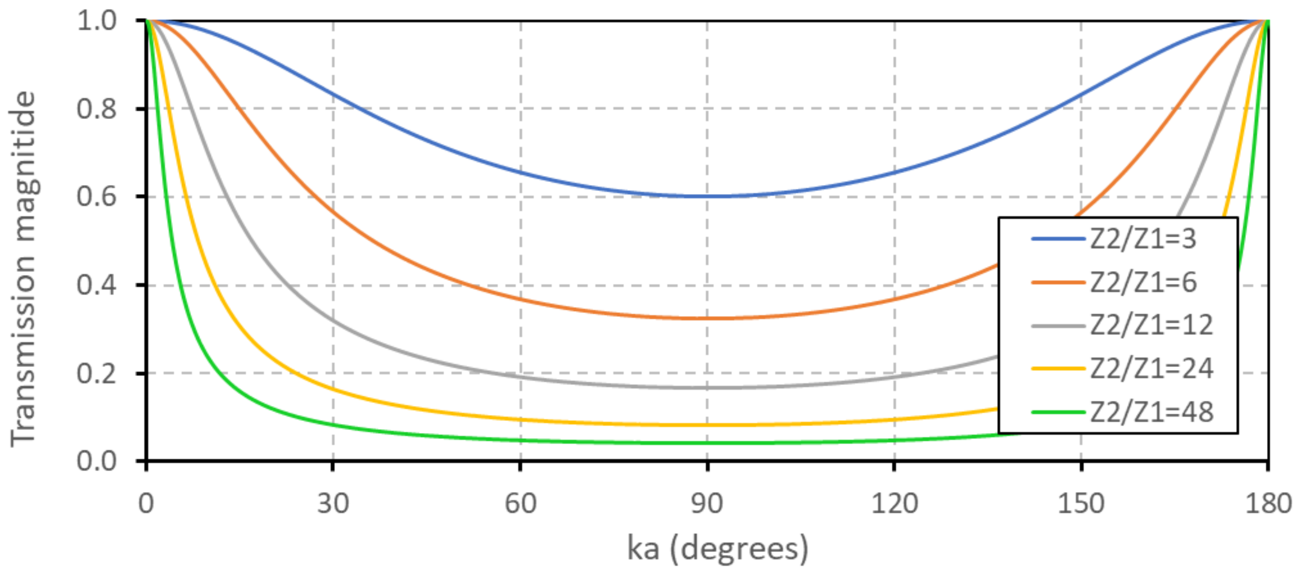
Solving gives $\quad F_3 = \dfrac{e^{jka}}{1 - A\left(e^{2jka} - 1\right)} F_1 \quad$ where $\quad A = \dfrac{(Z_2/Z_1 - 1)^2}{4 Z_2/Z_1}$

For impedance discontinuities where $Z_2/Z_1 < 2$ then $A < 0.125$ hence if the perfection is in this range and its length is much shorter than the wavelength then it could be negligible.

Transmission for impedance discontinuity



$$\left|\frac{F_3}{F_1}\right|$$



This transmission plot shows the higher impedance track giving a similar effect to an inductor.

For a small length a compared to the wavelength there is almost full transmission.

For lengths with $ka < \pi/2$ then for increasing frequency (increasing k) transmission is reduced.

Very high levels of attenuation is only possible if the impedance ratio Z_2/Z_1 is very large which makes the track extraordinary narrow hence this type of inductor is rarely used.

Transmission Line Stubs

A stub is a section of transmission line only connected to the signal line at one end, the other end is usually terminated with an open circuit or a short circuit.

Most line types including microstrip and grounded coplanar waveguide stubs can only be physically connected as a shunt between the signal and ground rather than as a series element.

From slide 6 dividing voltage by current for sine waves gives impedance along a line as

$$Z(z) = Z_o \frac{F\exp(jkz) + R\exp(-jkz)}{F\exp(jkz) - R\exp(-jkz)}$$

For a transmission line stub with intrinsic impedance Z_o terminated with load Z_L using (18.1) gives

$$R(Z_L + Z_o) = F(Z_L - Z_o)$$

Hence

$$Z(z) = Z_o \frac{(Z_o + Z_L)\exp(jkz) + (Z_L - Z_o)\exp(-jkz)}{(Z_o + Z_L)\exp(jkz) - (Z_L - Z_o)\exp(-jkz)} = Z_o \frac{Z_L\cos(kz) + jZ_o\sin(kz)}{Z_o\cos(kz) + jZ_L\sin(kz)}$$

i.e. the termination determines the forward and reflected waves F and R

Open

$Z_L = \infty$ $\qquad Z_{open} = -jZ_o\cot(ka)$

Short

$Z_L = 0$ $\qquad Z_{short} = jZ_o\tan(ka)$

For $ka < \pi/2$ the signs and k dependence imply that Z_{open} acts as a capacitor and Z_{short} acts as an inductor.

For stub length 1/8 of a wavelength $ka=\pi/4$ at the frequency of interest

$$Z_{open} = -jZ_o \qquad Z_{short} = jZ_o$$

Filters

In communication systems filters are used to separate channels, which when closely spaced the filter need a very sharp cut-off.

For accelerator systems the requirement is usually to remove well separated, unwanted frequencies generated by non-linear components such as mixers.

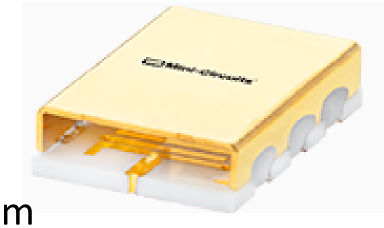
Huge numbers of filters are commercially available, so designing one's own is rarely necessary.

Responses are given in datasheets so the knowing the type (Butterworth, Elliptic, Chebyshev, Bessel and M-derived filters etc.) may not be important.

Often one wants a smooth and almost flat response in the frequency range of interest hence a Butterworth filter might be one's first choice.

Principal filter fabrications are lumped LC, LTCC (low temperature co-fired ceramic) ceramic, thin film, Surface Acoustic Wave (SAW), MMIC, PCB, tubular co-axial, waveguide and cavity.

Ceramic



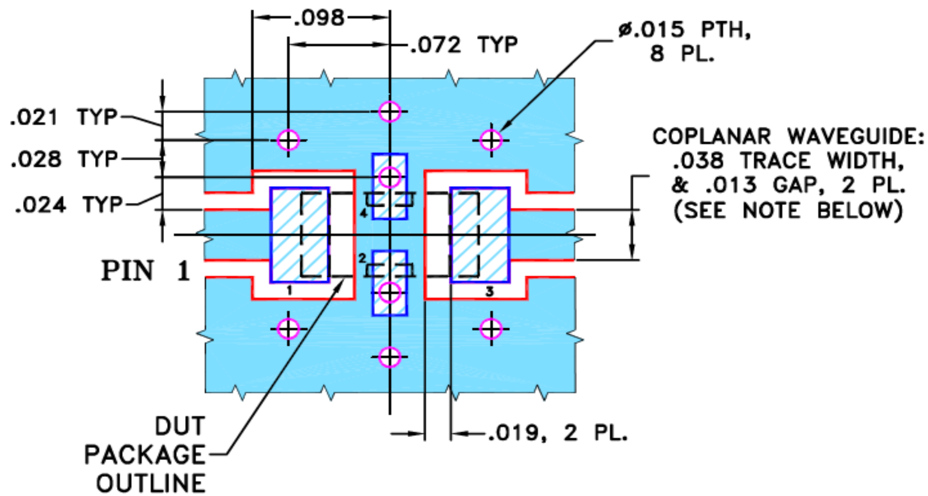
thin film

Tubular lumped element

Manufacturers have product selection tools focusing on final performance.

Mini-circuits LTCC BFCN-3010+ ~ 5 Euro

Mini-circuits supply Low temperature co-fired ceramic components throughout the frequency range of about 400 MHz to 50 GHz [7]. Note that thin film and MMIC filters are also available as surface mount and typically have very low reflection. For low frequencies < 400 MHz lumped LC filters are most commonly listed. The ceramic filters give the benefit of narrow passbands.

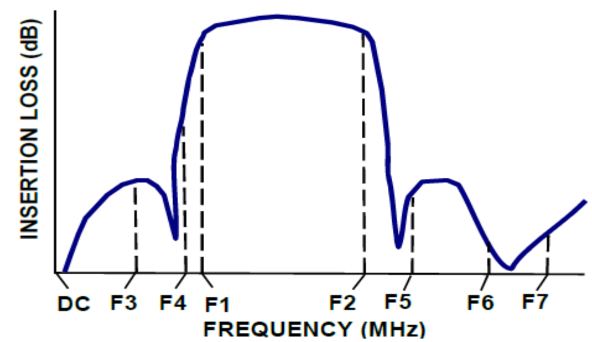


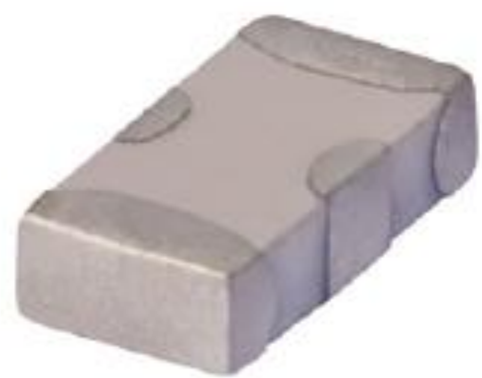
NOTES: 1. COPLANAR WAVEGUIDE PARAMETERS ARE SHOWN FOR ROGERS RO4350B WITH THICKNESS .020" ± .0015".
COPPER: 1/2 OZ. EACH SIDE.
FOR OTHER MATERIALS TRACE WIDTH & GAP MAY NEED TO BE MODIFIED.

2. BOTTOM SIDE OF THE PCB IS CONTINUOUS GROUND PLANE.

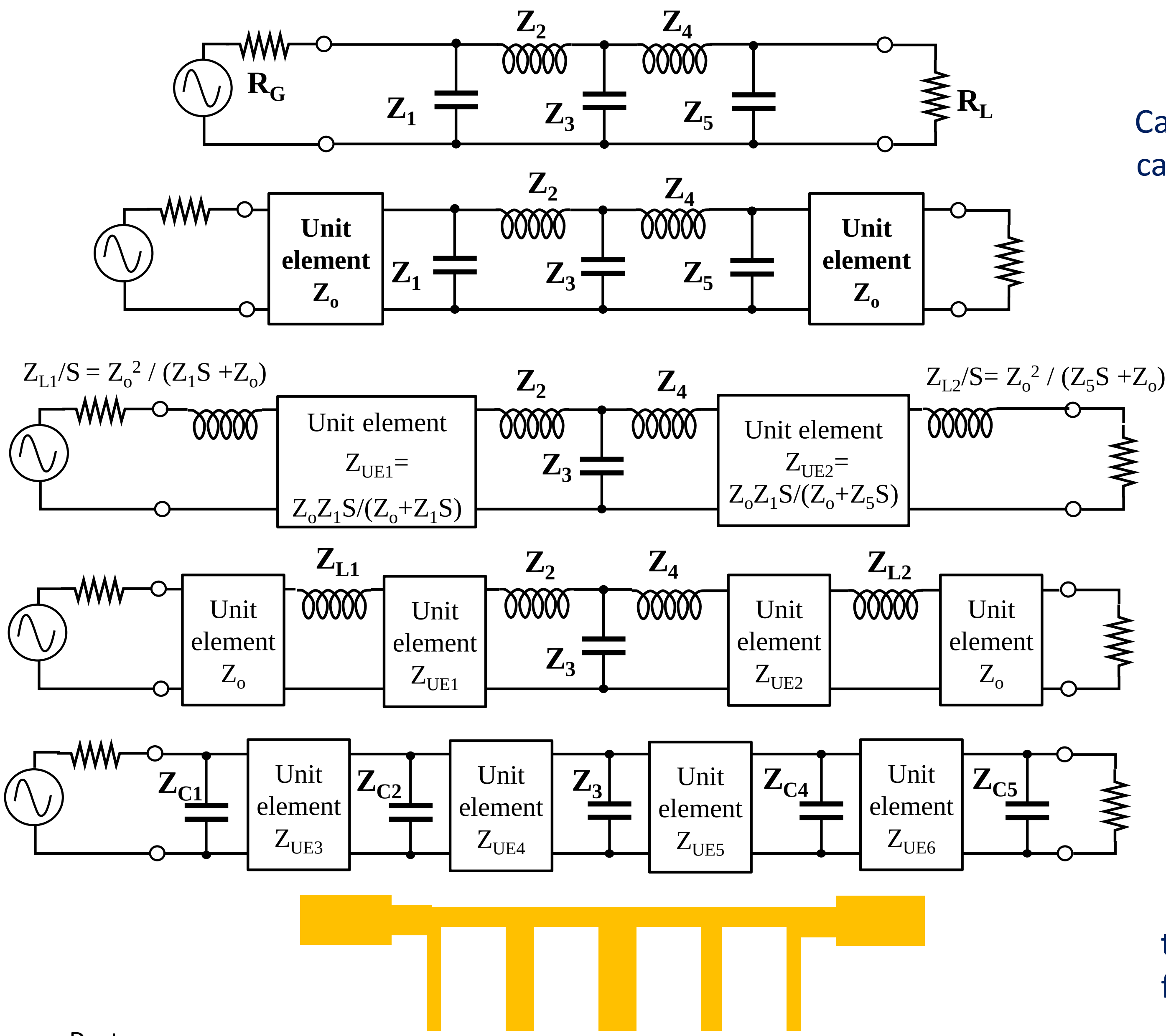
DENOTES PCB COPPER LAYOUT WITH SMOBC (SOLDER MASK OVER BARE COPPER)

DENOTES COPPER LAND PATTERN FREE OF SOLDER MASK





Printed Filters on PCBs



Capacitances to ground can be implemented as open shunt stubs.

Series inductors cannot be implemented

Solution uses Unit elements and Kuroda identities

A unit element is a quarter wavelength transmission line

The ladder is progressively transformed a form for implementation

Richards transformation

Filters designs start with a lumped circuit of capacitors and inductors.

Implementation requires lumped capacitors and inductors to be much smaller than the wavelength.

At high frequencies a distributed analysis is required.

A transmission line stub is a distributed circuit with a well-defined response and can act as a capacitor or an inductor for a limited range of frequencies.

The Richards transformation is used to map the response of a filter design realised with lumped elements to one with realised waveguide stubs.

The responses are only similar close to the frequency of interest. Richards chose stubs of length 1/8 of wavelength at the angular frequency of interest ω_o so that for other wave numbers k and associated frequencies ω one gets $ka = \tfrac{1}{4}\pi\omega/\omega_o$

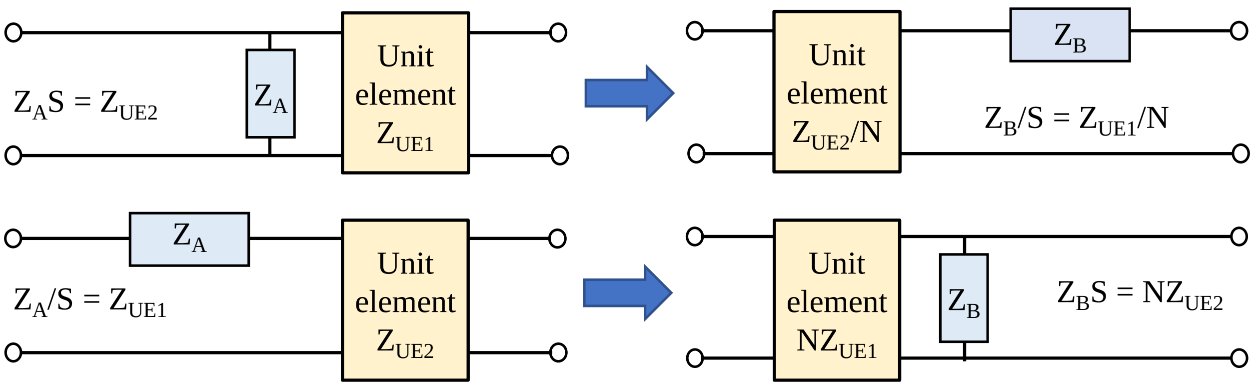
For an inductor the Richards transformation is $j\omega L \Rightarrow jZ_o \tan\left(\tfrac{1}{4}\pi\omega/\omega_o\right)$

For a capacitor the Richards transformation is $1/j\omega C \Rightarrow -jZ_o \cot\left(\tfrac{1}{4}\pi\omega/\omega_o\right)$

The Richards transformation maps the lumped element frequency response from the range $0 \le \omega < \infty$ to the range $0 \le \omega < 2\omega_o$

When using transmission line stubs, the filter response becomes periodic which unfortunately means there can be multiple pass bands when only one was desired. If the filter does not exclude all the unwanted frequencies it will need to be used with additional filters to block other ranges.

Two of Kuroda's identities



Kuroda identities can transform series components into shunt components.

$N = 1 + Z_{UE2}/Z_{UE1}$

Z_{UE} = intrinsic impedance of transmission line

They are easily proved by multiplying ABCD matrices

SERIES

$$\begin{bmatrix} v_{in} \\ i_{in} \end{bmatrix} = \begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix} \begin{bmatrix} v_{out} \\ i_{out} \end{bmatrix}$$

SHUNT

$$\begin{bmatrix} v_{in} \\ i_{in} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 1/Z & 1 \end{bmatrix} \begin{bmatrix} v_{out} \\ i_{out} \end{bmatrix}$$

Unit Element

$$\begin{bmatrix} v_{in} \\ i_{in} \end{bmatrix} = \begin{bmatrix} 1 & S Z_{UE} \\ S/Z_{UE} & 1 \end{bmatrix} \begin{bmatrix} v_{out} \\ i_{out} \end{bmatrix}$$

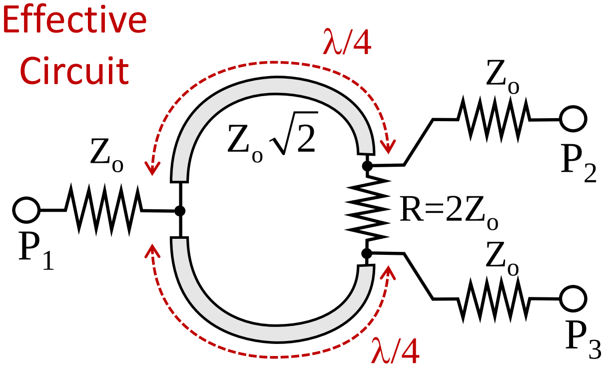
The identities work for any S but for a microstrip filter use the Richards transformation so that

$$S = j\tan\left(\tfrac{1}{4}\pi\omega/\omega_o\right)$$

If inductors are required as series components in top right and bottom left images, S must be complex and positive. When S is complex and positive, then shunt components must be capacitive as S multiplies Z_A and Z_B so would cancel a j in the denominator, for example $1/j\omega C$ to give a real positive value for Z_{UE}.

Wilkinson Splitter

Effective Circuit



The Wilkinson Splitter can also be used as a combiner

Excellent isolation between ports 2 and 3.

Splitting and combining is nominally lossless.

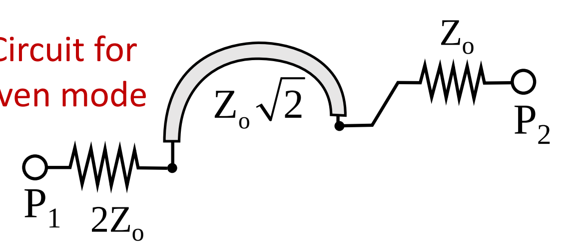
S Matrix

$$\frac{j}{\sqrt{2}}\begin{bmatrix}0 & 1 & 1\\ 1 & 0 & 0\\ 1 & 0 & 0\end{bmatrix}\begin{bmatrix}F_{input}\\ R_{out1}\\ R_{out2}\end{bmatrix} = \frac{j}{\sqrt{2}}\begin{bmatrix}R_{out1}+R_{out2}\\ F_{input}\\ F_{input}\end{bmatrix}$$

Splitter arms are quarter wave transformers.
Z_o is the impedance of input and output lines

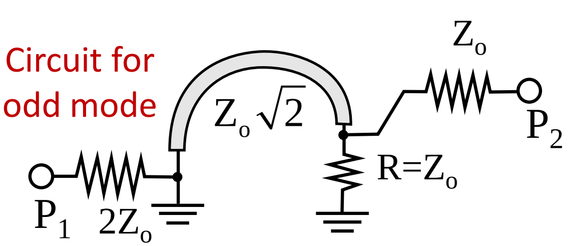
The routing of the quarter wave transformers is often chosen to minimise cross coupling, keep the resistor small and minimise reflection associated with track curvature.

Circuit for even mode



Solutions for linear circuits can be constructed by adding other simpler to determine solutions. For analysis of inputs to ports P_2 and P_3, odd and even excitations are considered.

Circuit for odd mode



Solutions must be constructed for inputs to each of the 3 ports. For input to port 1 the solution is symmetric hence no current in R. Only half the circuit needs to be drawn – hence the circuit for the even mode.

Even and odd mode analysis

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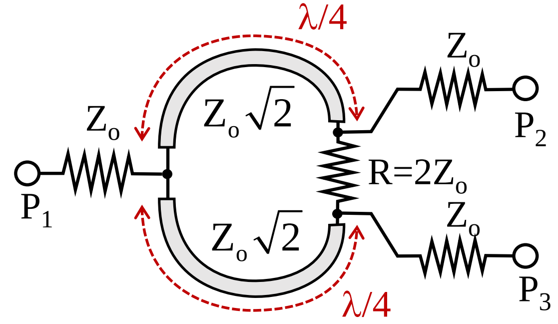
To construct an input to just P_2 or P_3 one adds the odd mode to the even mode, i.e.

$$P_2 = 0.5\,(P_2 + P_3) + 0.5\,(P_2 - P_3)$$

$$\frac{j}{\sqrt{2}}\begin{bmatrix}0 & 1 & 1\\ 1 & 0 & 0\\ 1 & 0 & 0\end{bmatrix}\begin{bmatrix}F_{input}\\ R_{out1}\\ R_{out2}\end{bmatrix} = \frac{j}{\sqrt{2}}\begin{bmatrix}R_{out1}+R_{out2}\\ F_{input}\\ F_{input}\end{bmatrix}$$

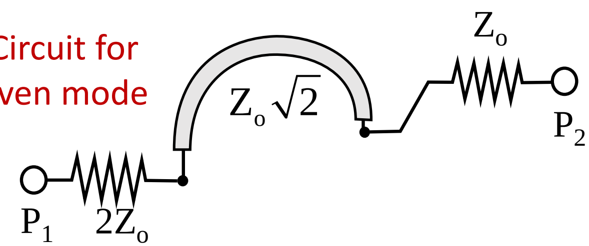
Even mode input impedance $2Z_o$ is perfectly matched to output impedance Z_o with a quarter wave transformer. Solution gives no reflection at port P_1.



As the S matrix is reciprocal then one also deduces that equal inputs at P_2 and P_3 add together at P_1 as an output.

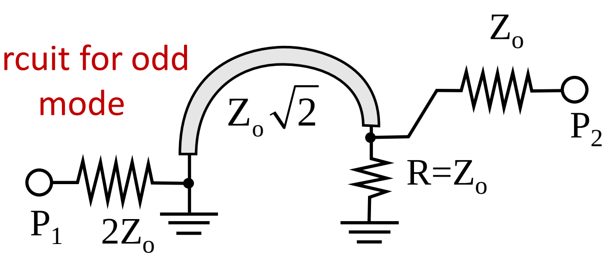
To complete the analysis odd mode excitation must be determined. For odd mode excitation the centre line must become an earth plane as shown.

Circuit for even mode



For odd mode input P_2 sees resistance of Z_o to earth shunted with a shorted quarter wave transformer.

Circuit for odd mode



A a shorted quarter wave transformer has infinite impedance hence the odd mode is perfectly matched with all the power being dissipated in the resistor.

The S matrix has now been fully determined.

Wilkinson Splitters on phase board

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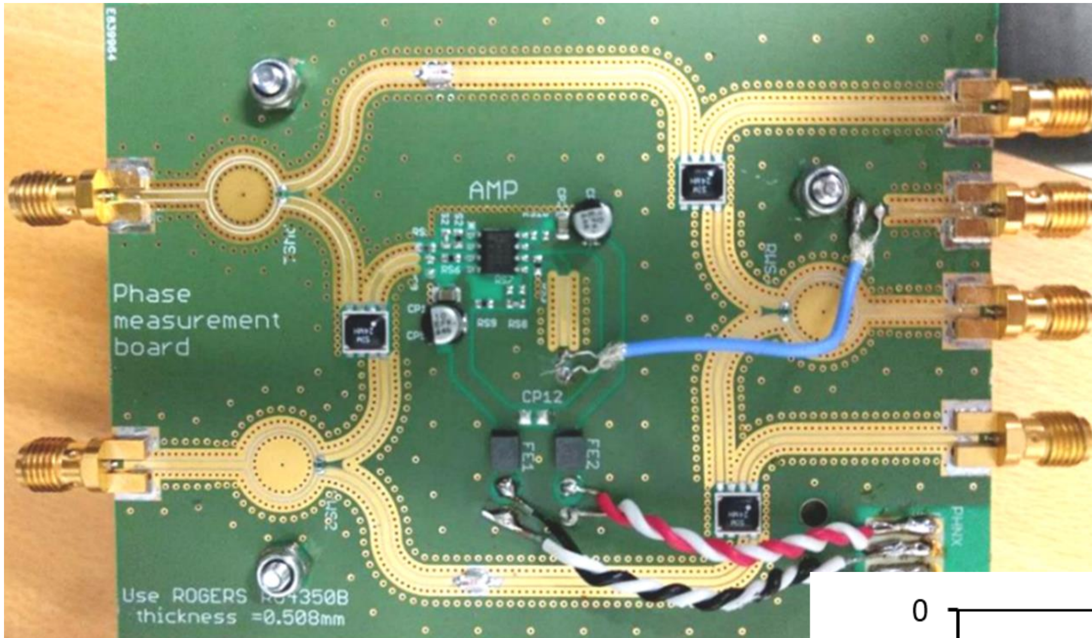
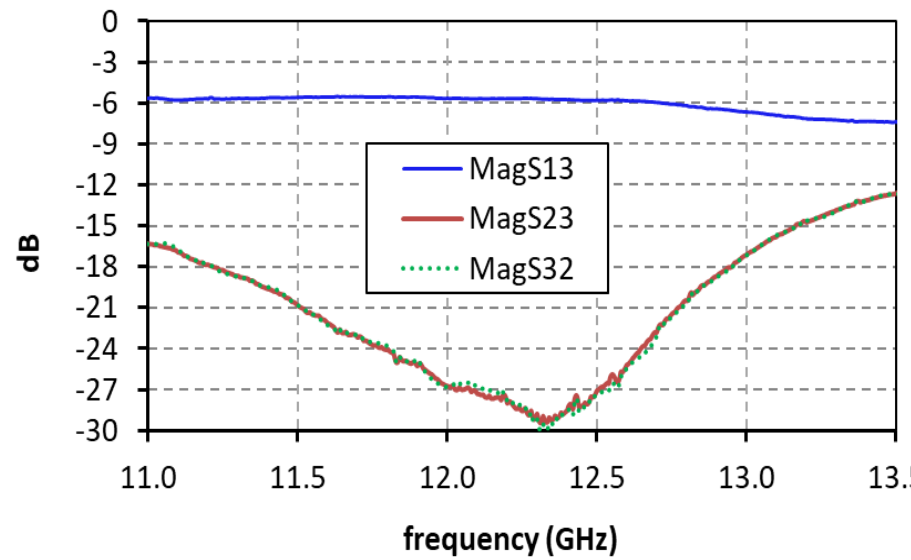


The Wilkinson splitters on the phase board use ¾ wavelength transformers. The measured S parameters below are for the test board. Only -27 dB was achieved at 12 GHz however modifications were made which should have moved the best isolation to 12 GHz.

The tracks on the phase board are a precise number of wavelengths hence the effect of reflection which could have been -30 dB is further reduced possibly by a similar factor.

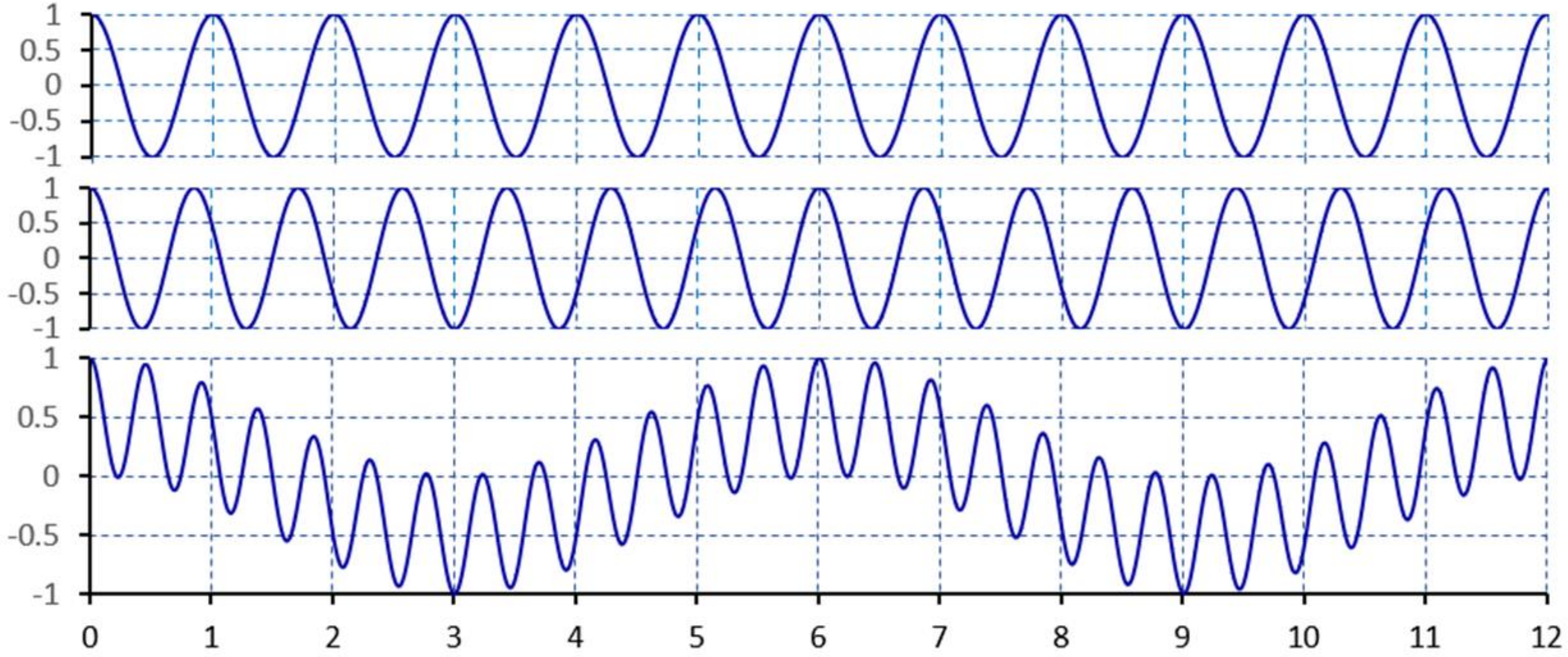


Mixing

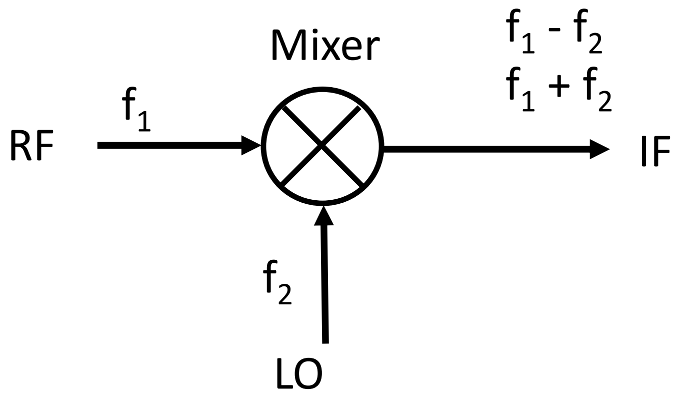
- Mixers perform time domain multiplication of signals.
- For perfect multiplication the output only contains the sum and difference frequencies.

$$2\cos(\omega_1 t+\theta_1)\cos(\omega_2 t+\theta_2)=\cos\{(\omega_1+\omega_2)t+\theta_1+\theta_2\}+\cos\{(\omega_1-\omega_2)t+\theta_1-\theta_2\}$$

- Output phases are also sums and differences of the input phases.
- Output with the summed frequency is known as the up converted signal.
- Output with the difference frequency is known as the down converted signal.



Mixers



For a communication system,

The local oscillator (LO) is sinusoidal

The RF carries information to be transmitted by amplitude and phase modulation of f_1

If f_1 is close to f_2 then the output $f_1 - f_2$ will be at a much lower frequency for convenient processing and is called the intermediate frequency (IF)

Low pass and high pass filter select the output

It is not possible to perform a perfect multiplication with simple analogue devices. For a real mixer only part of the output is a pure multiplication

Types

Operation

Add and square, multiply by square wave

Input and Output Connection

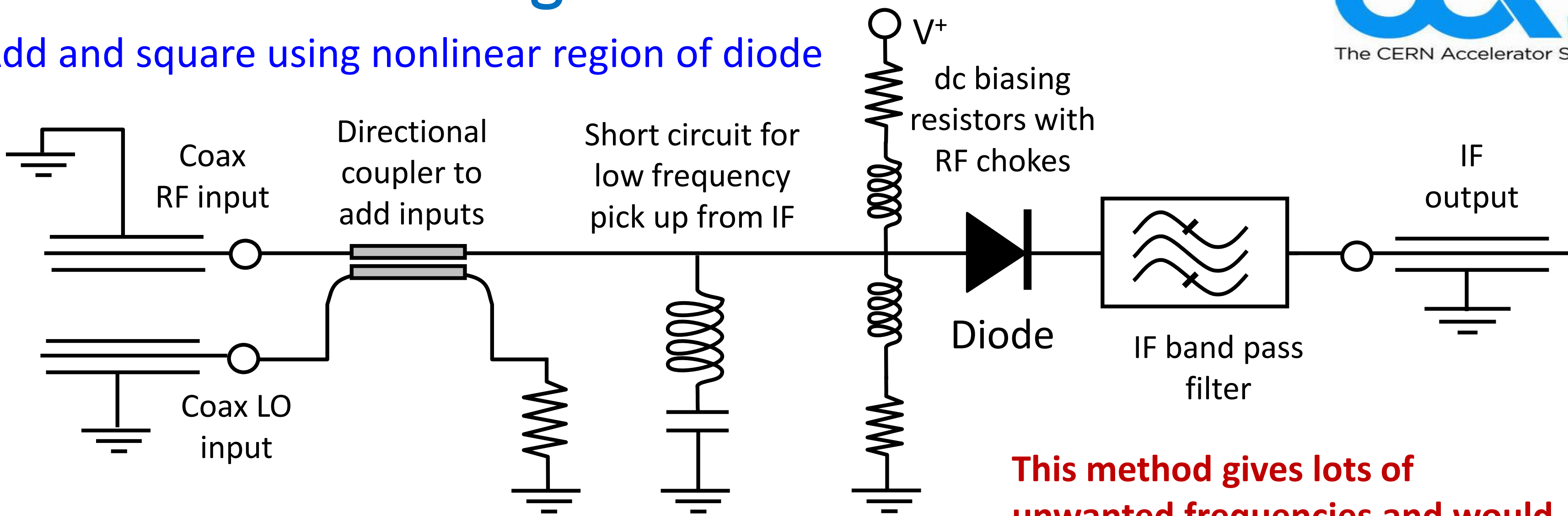
Unbalanced - Single balanced – Double balanced – Triple Balanced, Gilbert Cell

Devices

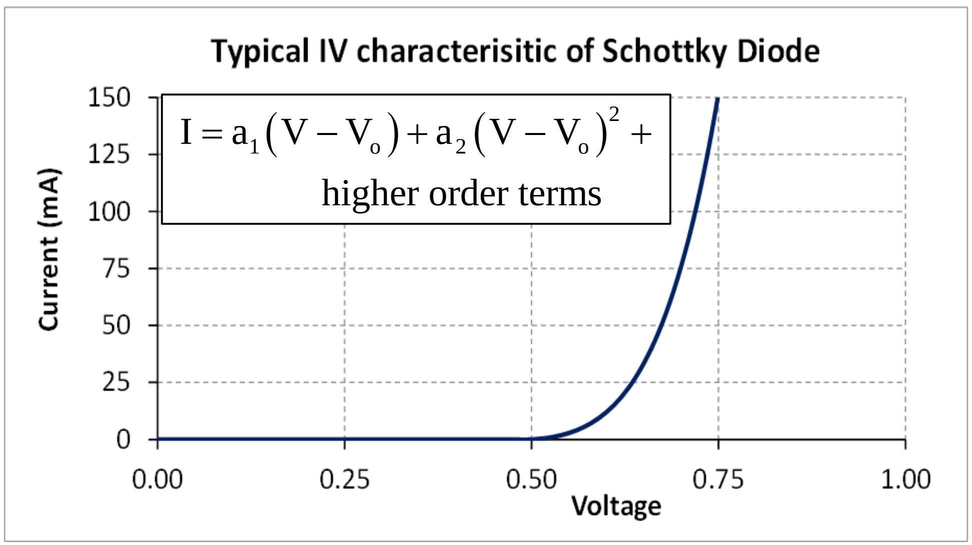
PN junction diodes, Schottky diodes, BJTs, FETs

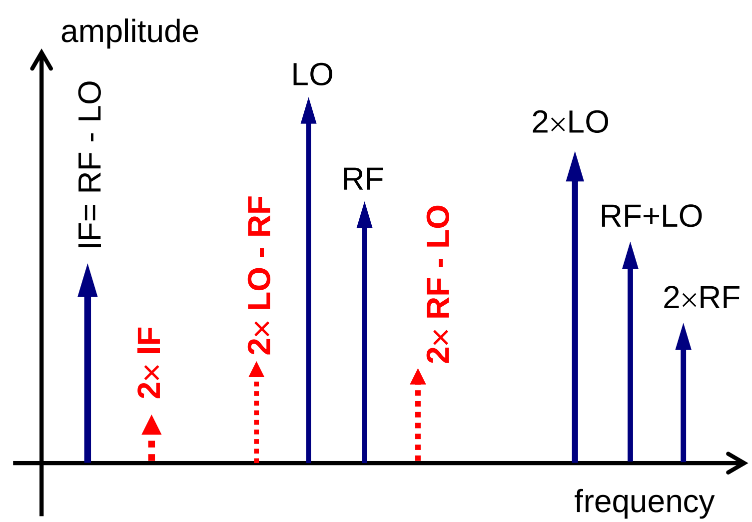
Unbalanced Mixing

Add and square using nonlinear region of diode



This method gives lots of unwanted frequencies and would not be used for accelerators





Higher order terms

For an arbitrary device where V is the input

$$V_{out} = a_o(V_o) + a_1 V + a_2 V^2 + a_3 V^3 + \cdots + a_k V^k + \cdots$$

For input V containing two frequencies

$$V = A_1 \cos 2\pi f_1 t + A_2 \cos 2\pi f_2 t$$

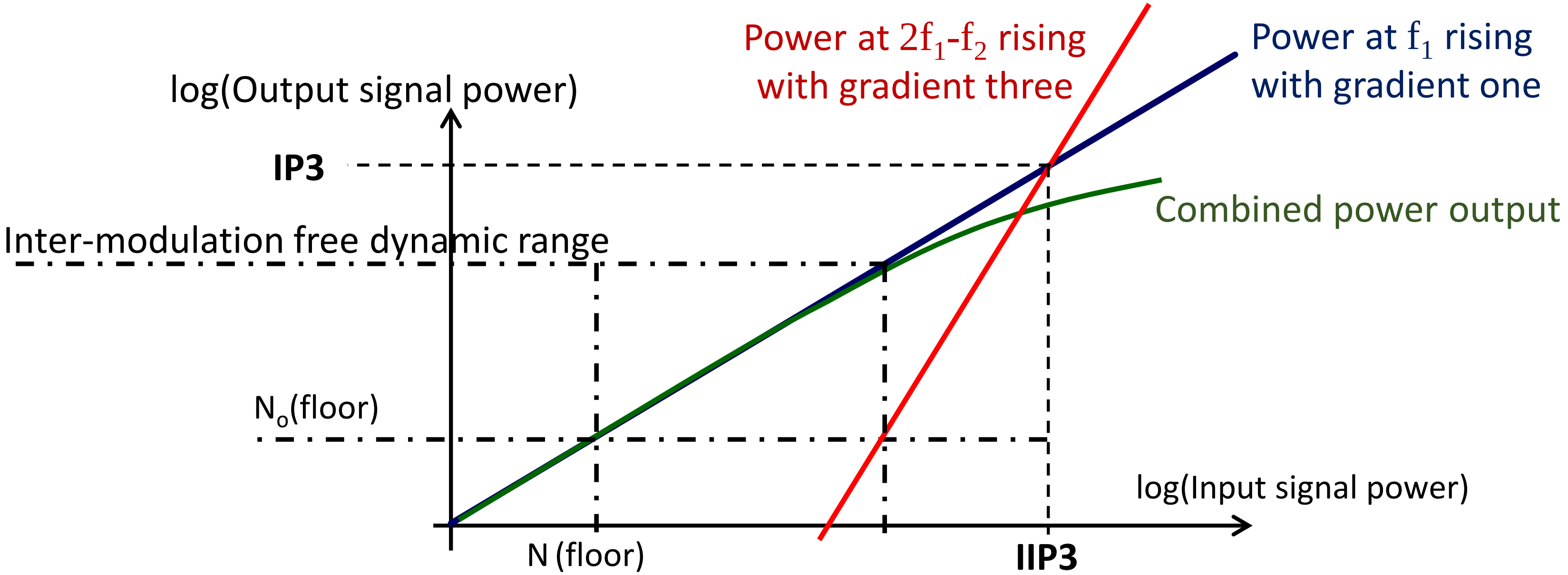
Output frequencies determined as

$$f_{out} = |m f_1 + n f_2|$$

Where f_{out} is positive and m and n are positive or negative integers

Frequencies from k[th] harmonic determined with additional constraint $|m| + |n| = k$

Nonlinearities can map odd harmonics back near the fundamental



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