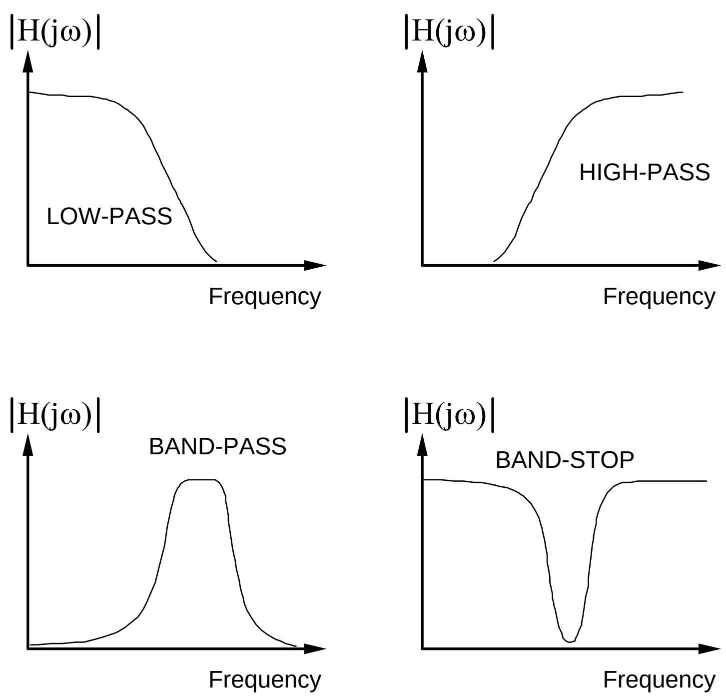
# RF & Microwave Filters: Lumped & Distributed

From the introduction to radio transmitters and receivers it is clear that filters are a very crucial component for separating wanted signals from unwanted signals of other users and systems. This section first gives an introduction to the theory, design procedure and practical construction of RF filters based on inductors and capacitors. The purpose of most linear filters is to separate a wanted signal from a mixture of the wanted signal and one or more unwanted signals. A filter may be defined as: "a transducer for separating waves on the basis of their frequencies".

**Filter that are required can be:-**

- Low-pass
- High-pass
- Band-pass
- Band-stop
- All-pass (but with some specific phase response)





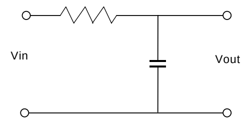

**Passbands**: The frequency band occupied by the wanted signal is referred to as the *passband*. In this band the ideal requirement for the filter is to provide constant loss so that the wanted signal will be transmitted with no distortion. With a finite, lumped network one cannot achieve exactly constant loss over a finite band of frequencies, so it is customary to specify instead some acceptable upper and lower limits between which the loss can vary.

**Stopbands**: The frequency bands occupied by the unwanted signals are referred to as the *stopbands*. As far as the filter user is concerned, it is normally of no interest whether the actual filter loss in the stopband exceeds the amount specified by 1 dB or 20 dB, and in this sense stopband responses have only lower limits in contrast to passbands where there are both upper and lower limits.

**Transition bands**: The loss response of a filter is a continuous function of frequency and is thus unable to have sudden discontinuities. For this reason there must always be some interval in the frequency spectrum, separating the edge of the passband from the edge of the stopband, in which the loss can rise from the low value in the passband to that required in the stopband. These intervals are described as *transition bands*.

## TRANSFER FUNCTION ANALYSIS

To start to understand the deep complexities of filter theory, it is necessary to briefly study transfer functions and Bode plots. Consider the simple R-C lowpass filter shown.



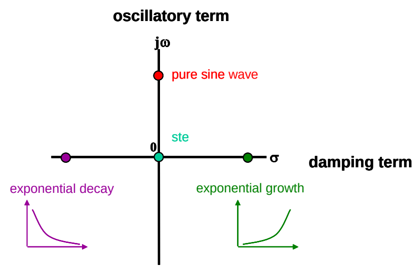

Using the ordinary potential divider expression, we can easily find that:-

$$V_{out} = \frac{1/j\omega C}{R + 1/j\omega C} V_{in} = \frac{1}{1 + j\omega CR} V_{in}$$

(note that this assumes that assumes zero source impedance and infinite load impedance)

This transfer function represents the frequency-domain response of the network. The circuit is a lowpass filter and when $\omega = 1/RC$, the $|\text{voltage}|$ is divided by $\sqrt{2}$ (the 3dB point).

By using this analysis we have assumed that the signal is a pure sinusoid. In order to represent the filter more generally, the "jω" terms are all replaced by the complex "s" operator ($s=\sigma+j\omega$). The "s" operator is able to represent a wide range of signals, such as step functions, impulses, and sinusoids:-



**For the same RC lowpass filter, we write the transfer function as:-**

$$H(s)=\frac{1/sC}{R+1/sC}=\frac{1}{1+sCR}$$

The transfer function H(s) is related to the time-domain response of the network by the LAPLACE TRANSFORM.

The IMPUSE RESPONSE is the output, r (t), of the network when an ideal unit impulse δ(t) is the input signal:

$$r(t) = L^{-1}[H(s)]$$

For a general network the transfer function will have a polynomial for both the numerator *N (s)* and for the denominator *D (s)*:-

$$H(s)=\frac{a_p s^p + a_{p-1}s^{p-1}+\ldots+a_1 s + a_0}{b_q s^q + b_{q-1}s^{q-1}+\ldots+b_1 s+b_0}$$

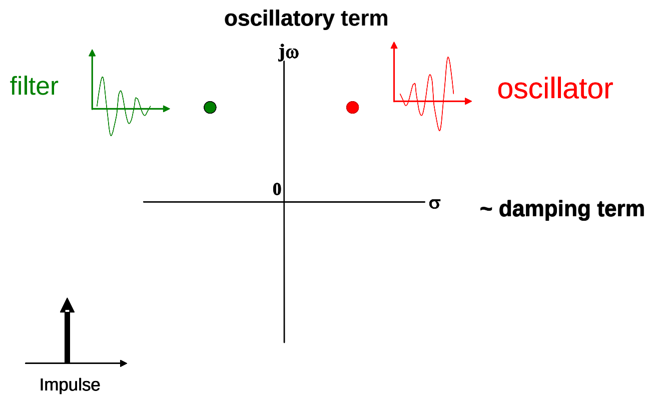
of course, these polynomials can be expressed in terms of a product of their factors:-

$$H(s)=K.\frac{(s+z_1)(s+z_2)\ldots(s+z_n)\ldots}{(s+p_1)(s+p_2)\ldots(s+p_n)\ldots}.$$

The quantities $-z_1$, $-z_2$, … are called the zeros, as the transfer function becomes zero at all these values.

The quantities $-p_1$, $-p_2$, … are called the poles, as the transfer function goes to infinity at these values.

The poles and zeros can be anywhere in the COMPLEX *s*-domain:-

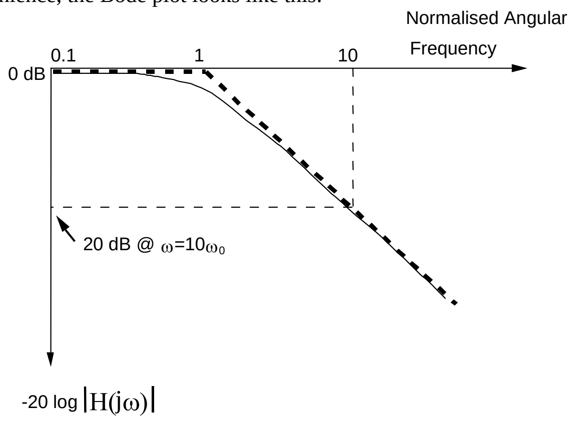





**BODE PLOTS**

**The Bode plot is graphical technique for representing transfer functions (both magnitude and phase) vs. frequency. Each pole and zero can be approximated by a straight line when plotted on a log-log scale.**

For example, for the previous simple lowpass filter with $H(s)=\frac{1}{1+sCR}$, and by normalising to CR=1 for convenience, the Bode plot looks like this:-



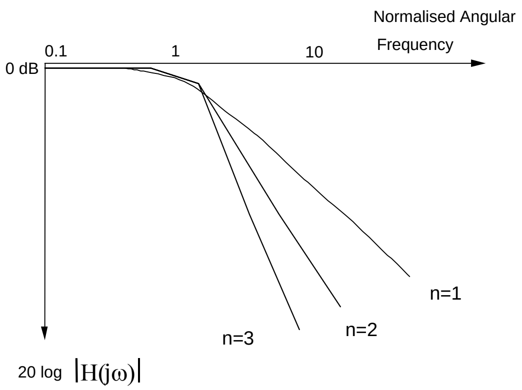

Each pole and zero is represented by a straight line with slope "20dB per decade", either up or downwards. So, you can imagine that by deliberately putting poles and zeros into the response at certain places, almost any desired frequency response can be obtained.

For example, the well-known Butterworth low-pass filter has a normalised transfer function given by:-

$$H(s) = \frac{1}{(1+s^n)}$$

$$|H(j\omega)|=\frac{1}{\sqrt{(1+\omega^{2n})}}$$

At the normalised cut-off frequency $\omega$=1, clearly the magnitude of $H(j\omega)$ is 1/√2 (the -3 dB point). The value of '*n*' determines how many poles are operating, and therefore the steepness of the roll-off.




**Butterworth filter: Achieving a sharper cut off by increasing the order**

"Poles" in filters can be a source of considerable confusion because they are on the DENOMINATOR, but when the signal frequency approaches a pole, the filter's transfer function starts to **fall**. It is very important to remember that the pole is in the complex S-domain. There is no signal generator in the lab that can generate this signal. Real world CW test signals are on the vertical axis in the S-domain. What matters then is how close the pole is to that vertical axis. With incorrect design of the lowpass filter, with a pole too close to the vertical axis, a very sharp peak in the frequency response is created. With proper design, the desired smooth roll-off is obtained:-

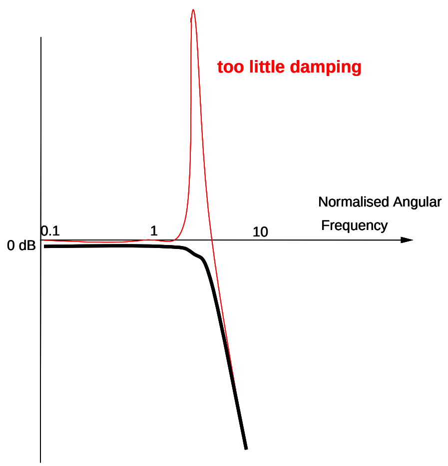


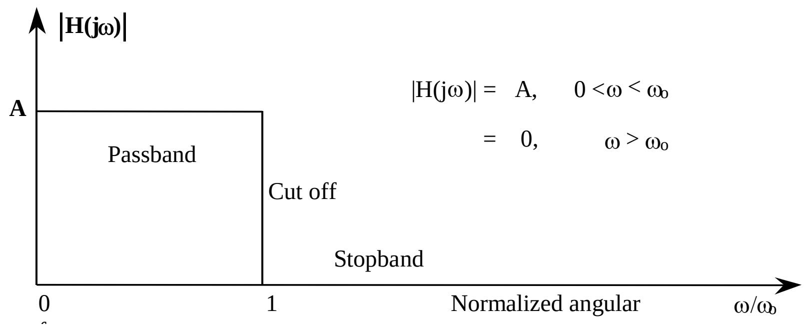

**Illustration of how with proper design a POLE leads to a filter response that falls.**



## FILTER THEORY

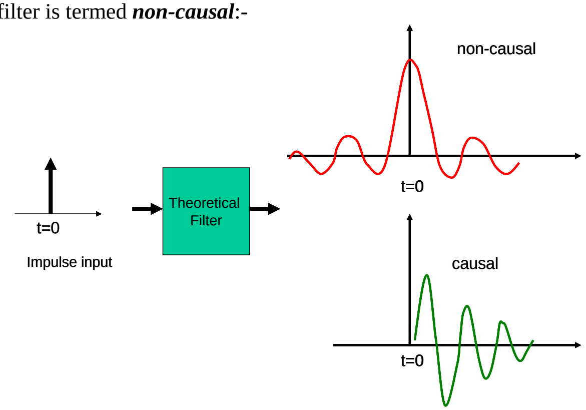
Ideal characteristics such as perfectly sharp cut-offs, infinite attenuation in the stopband, and zero attenuation in the passband (i.e. so-called 'brick wall' characteristics) cannot be realised in a practical circuit. These ideal characteristics for a lowpass filter are defined as:

$$|H(j\omega)| = A, \quad 0 < \omega < \omega_0$$
$$= 0, \quad \omega > \omega_0$$



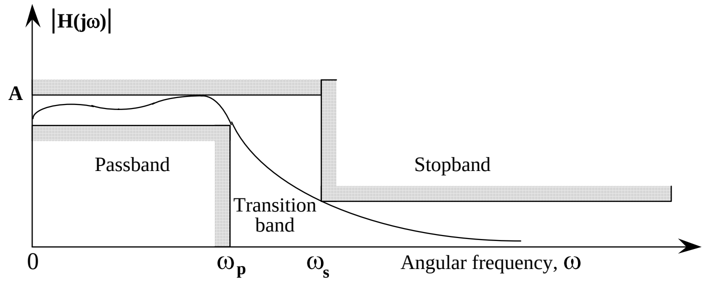

**Ideal lowpass filter response**

Unfortunately, the inverse Laplace transform of this brick wall response, which gives the time-domain impulse response, shows that there is an output signal **before** the impulse at $t$=0 is even applied. Such a filter is termed ***non-causal***:-




In the digital signal processing world, this problem can be partly tackled by putting a delay into the filter. DSP can do this by using digital memory devices. Delays are OK in digital systems, up to a point. In digital TV, for example, you may be familiar with pausing live TV, to make a cup of tea: The incoming digital data is stored onto a hard disk, whilst you are away, and then when you come back it is read back and you watch from where you left off. Whilst you are watching the "old signal" from (say) 10 minutes ago, the incoming new TV signal is being stored to disk in the background. In analogue systems, information storage has no doubt been attempted but is basically impractical.

So, in reality in the analogue world, we may only approximate the ideal brick wall filter characteristics within a certain prescribed error. Thus a typical realisable approximation to the ideal brick-wall filter is illustrated. Here the response is constrained to lie between the shaded areas in the prescribed passband and stopband. We shall consider three filter types for approximating the ideal filter characteristics, namely, 'Butterworth' (or maximally flat), 'Chebyshev' (or equi-ripple), and 'elliptic'. Bandpass, bandstop, and highpass filter networks can be derived from a lowpass filter network, so it is useful to consider first how the lowpass characteristics are specified.



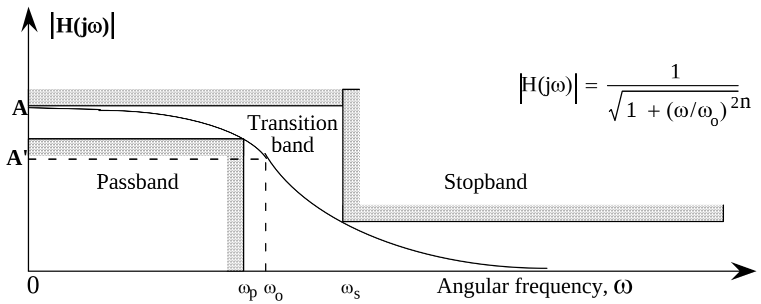

**Realistic lowpass filter response**

*Butterworth filter*

The amplitude response of the Butterworth filter is given by:

$$\left|H(j\omega)\right| = \frac{1}{\sqrt{1 + (\omega/\omega_o)^{2n}}}$$

The frequency response of this filter is illustrated below. Note the flatness of the response in the passband, with no ripples present.




**Butterworth (or Maximally Flat) Filter Response**



The attenuation $\alpha_n(\omega)$ as a function of angular frequency is specified as:

$$\alpha_n(\omega) = 10\log\{1 + (\omega/\omega_o)^{2n}\}$$

where $\omega_o$ is the frequency for which $\alpha$=3dB ('half-power point'). $\alpha_{max}$ is the maximum attenuation allowed in the passband. $\alpha_{min}$ is the minimum attenuation allowed in the stopband. Hence,

$$\alpha_{max} = 10\log\{1 + (\omega_p/\omega_o)^{2n}\}$$

and

$$\alpha_{min} = 10\log\{1 + (\omega_s/\omega_o)^{2n}\}$$

The transfer function H(jω) has $n$ poles. To find them, consider the function $|H(j\omega)|^2$, and replace jω by the complex frequency $s$ . Since $|H(j\omega)|^2 = H(j\omega)H(-j\omega)$, replacing jω by $s$ results in H(s)H(-s). If $s_i$ is a pole of H(s) lying in the left half-plane, then $-s_i$, is a pole of H(-s), which is its mirror image in the right half-plane. The function H(s)H(-s) has 2$n$ poles of which $n$ lie in the left half-plane and give rise to a stable filter. Hence the desired poles are the values of $s$ lying in the left half-plane. Hence, the function

$$H(s)H(-s) = \frac{1}{1 + (-1)^n (s/\omega_o)^{2n}}$$

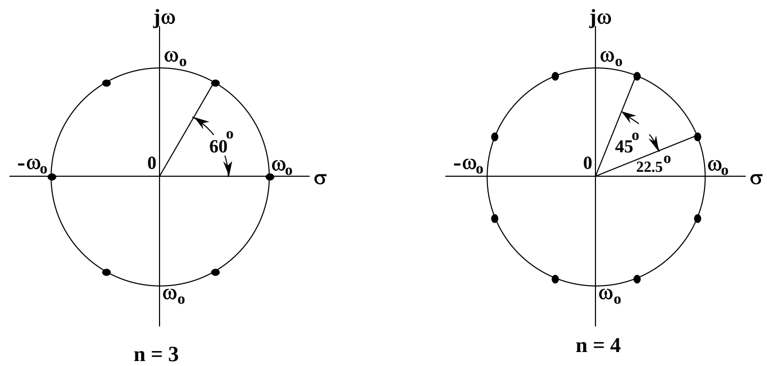
goes to infinity if $s$ is one of the 2n roots of the equations:

$$(s/\omega_o)^{2n} = +1, \text{ for } n \text{ odd},$$
$$= -1, \text{ for } n \text{ even}.$$

The roots are

$$s_i = \omega_o e^{j\pi(2i/n)/2} \qquad \text{for } n \text{ odd},$$
$$s_i = \omega_o e^{j\pi((2i-1)/n)/2} \qquad \text{for } n \text{ even}, \quad i = 1,...,2n.$$

These roots are equally spaced points on a circle of radius $\omega_o$. The poles of interest are the roots lying in the left half-plane.




**Poles in the s-domain for a Butterworth filters of order 3 and 4**

*Chebyshev filter*

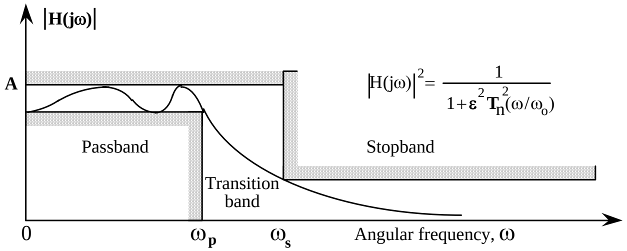
By allowing the magnitude of the filter response to have small ripples, one can achieve a sharper transition between the passband and the stopband. Here the lowpass characteristic may be approximated in the passband by an equi-ripple response. This equi-ripple passband and maximally flat stopband response leads to the *Chebyshev* filter. The magnitude of the passband varies between two prescribed limits, and then drops monotonically outside the passband. The magnitude of the transfer function is given by:

$$|H(j\omega)|^2 = \frac{1}{1+\varepsilon^2 T_n^2(\omega/\omega_o)}$$

where ε is a constant, and $T_n(\omega/\omega_o)$ is the Chebyshev polynomial, defined by:

$$T_n(\omega/\omega_o) = \mathrm{Cos}[n\mathrm{Cos}^{-1}(\omega/\omega_o)] \qquad \text{for } |(\omega/\omega_o)| < 1$$

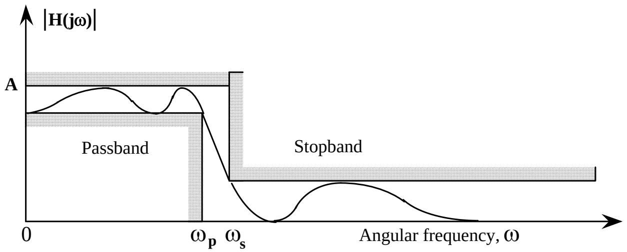
Let $\mathrm{Cos}^{-1}(\omega/\omega_o) = \theta$. Hence, this $T_n(\omega/\omega_o)$ is a polynomial because one can express Cos(nθ) as a polynomial in Cos(θ). For example, $\mathrm{Cos}(2\theta) = 2\mathrm{Cos}^2(\theta) - 1$, and $\mathrm{Cos}(3\theta) = 4\mathrm{Cos}^3(\theta) - 3\mathrm{Cos}(\theta)$, etc. Observe that $-1 \le T_n(\omega/\omega_o) \le 1$ for $|(\omega/\omega_o)| \le 1$ and therefore $|H(j\omega)|^2$ oscillates between $(1 + \varepsilon^2)^{-1}$ and 1 in the passband. At the cut-off frequency $(\omega/\omega_o) = 1$ and $T_n(\omega/\omega_o) = 1$. In terms of the poles in the s-domain, for a Chebyshev filter they are on an ellipse.




**Chebyshev Filter Response**

*Elliptic filter*

The elliptic filter exhibits equi-ripple behaviour in both the passband and the stopband. The characteristic behaviour provides a more rapid transition from passband to stopband than does the Chebyshev or Butterworth filter, but at the expense of a finite attenuation level in the stopband.

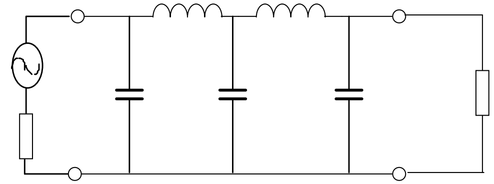



**Elliptical Filter Response**

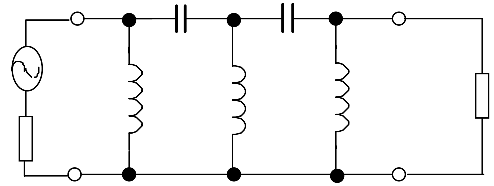


## PRACTICAL L-C FILTER DESIGN

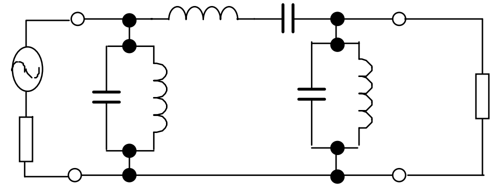
The filter circuits studied here are as follows:-



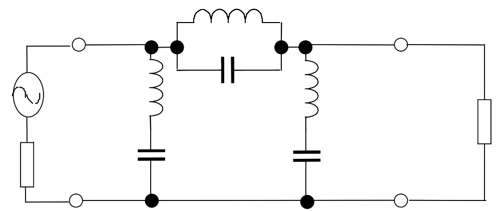
**LOW-PASS**



**HIGH PASS**

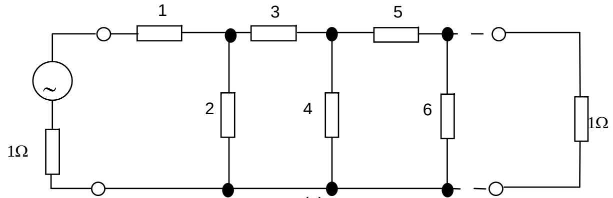


**BAND PASS**



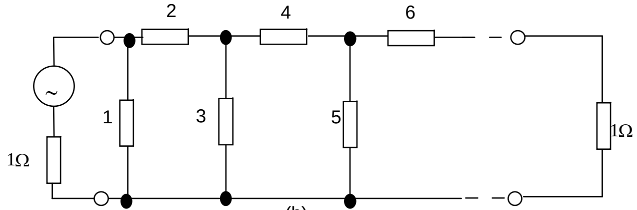
**BAND STOP**

Filter designs usually start with a low-pass filter which realises the required Butterworth/Chebyshev/Elliptic transfer function, with the component values normalised to an angular frequency of 1 rad/s and impedance of 1 Ω for convenience. This filter is called the “prototype” filter:-



(a)

OR



(b)

In the prototype filter the series elements represent inductors and the shunt elements represent capacitors. The normalised element values can be obtained from standard tables. For Butterworth filters the number of elements (the order of the filter) is the only parameter to be selected. For the other types there are also parameters of maximum ripple and insertion loss to be decided. The table for Butterworth prototype filters is:

| n | 1 | 2 | Element no. 3 | 4 | 5 | 6 |
|---|---|---|---|---|---|---|
| 2 | 1.414 | 1.414 | | | | |
| 3 | 1.000 | 2.000 | 1.000 | | | |
| 4 | 0.7654 | 1.848 | 1.848 | 0.7654 | | |
| 5 | 0.6180 | 1.618 | 2.000 | 1.618 | 0.6180 | |
| 6 | 0.5176 | 1.414 | 1.932 | 1.932 | 1.414 | 0.5176 |

For high-pass, band-pass, and band-stop filters the “prototype” low-pass filter has each of its elements transformed into a new element, in order to achieve the desired response.

## HIGHPASS TRANSFORMATION

$$\omega \rightarrow \frac{1}{\omega}$$

This means that the series inductors, with normalised value $g_k$, are replaced with shunt capacitors of normalised value $1/g_k$.



The shunt capacitors, with normalised value $g_k$, are replaced with series inductors of normalised value $1/g_k$.

Having carried out the transformation, the component values are “de-normalised”: Their values are simply scaled according to the cut-off or centre frequency required and the impedance of the generator and load.

Remembering that $Z_c = 1/j\omega C$ and $Z_l = j\omega L$, denormalising for lowpass and highpass is achieved as follows:-

$$L_k = \frac{normalised\ value_k \times R}{\omega_c} \qquad C_k = \frac{normalised\ value_k}{\omega_c R}$$

Where R is the source and load resistance. The $k^{th}$ normalised value is calculated by reading the *g*-values from the table, and then applying the transformation shown.

## BANDPASS TRANSFORMATION

For the bandpass, with upper and lower cut-off frequencies $\omega_2$ and $\omega_1$, respectively, the following frequency transformation is used:-

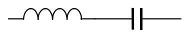
$$\omega \rightarrow \frac{1}{\Delta}\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right), \quad \text{where } \Delta = \frac{\omega_2 - \omega_1}{\omega_o} \quad \text{and} \quad \omega_o = \sqrt{\omega_1 \omega_2}$$

Note that $\omega_o$ is the *geometric* mean of the upper and lower angular cut-off frequencies; this is used instead of the arithmetic centre frequency for convenience in the mathematics.

Δ is the fractional bandwidth, and becomes an additional term in the transformation.

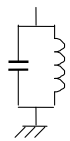
With the frequency and impedance scaling, the element values for the **series** resonators are then calculated from:-

$$L_k = \frac{g_k R}{\Delta \times \omega_o} \qquad C_k = \frac{\Delta}{\omega_o g_k R}$$



The element values for the **parallel** resonators are calculated from:-

$$L_k = \frac{\Delta R}{\omega_o g_k} \qquad C_k = \frac{g_k}{\Delta \times \omega_o R}$$



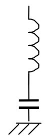
## BANDSTOP TRANSFORMATION

For a bandstop filter, the transformation is:-

$$\omega \to \frac{\Delta}{\left(\dfrac{\omega}{\omega_o} - \dfrac{\omega_o}{\omega}\right)}$$

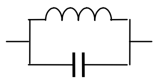
The element values for the shunt-connected **series** resonators are then calculated from:-

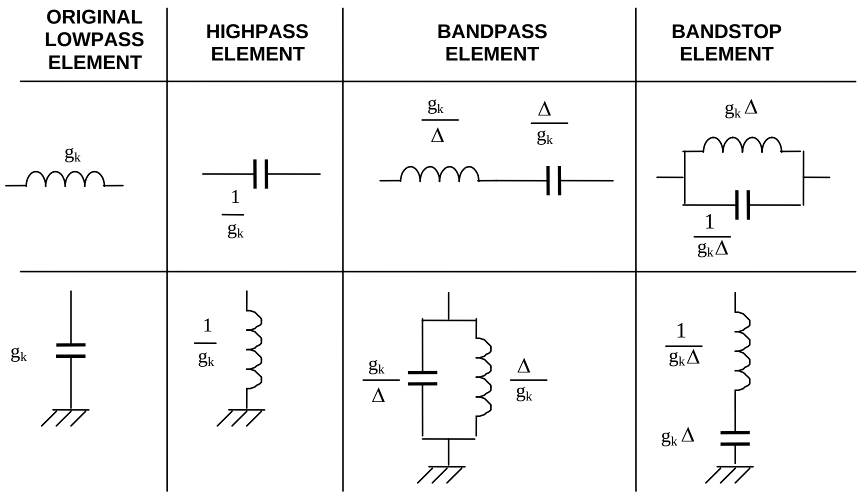
$$L_k = \frac{R}{\omega_o g_k \Delta} \qquad C_k = \frac{\Delta \times g_k}{\omega_o\, R}$$



The element values for the series-connected **parallel** resonators are calculated from:-

$$L_k = \frac{\Delta\, g_k R}{\omega_o} \qquad C_k = \frac{1}{\Delta \times \omega_o\, g_k R}$$



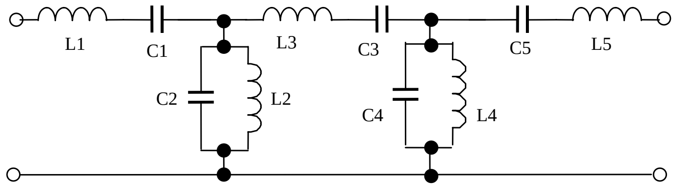



**SUMMARY OF FILTER TRANSFORMATIONS**



## BANDPASS FILTER DESIGN EXAMPLE

Q: Design a 5th-order Butterworth bandpass filter, with a passband from 65 to 75 MHz and 50 Ohm source and load impedances.




A: The lowpass prototype values are 0.618 1.618 2 1.618 0.618

$$\Delta = \frac{\omega_2 - \omega_1}{\omega_o} = \frac{f_2 - f_1}{\sqrt{f_1 f_2}} = 0.1432 \qquad \omega_o = 438699539$$

So, $$L_1 = \frac{0.618x50}{0.1432x438699539} = 492\mathbf{nH}$$

$$C_1 = \frac{0.1432}{0.618x50x438699539} = 10.6\mathbf{pF}$$

$$L_2 = \frac{0.1432x50}{1.618x438699539} = 10.1\mathbf{nH}$$

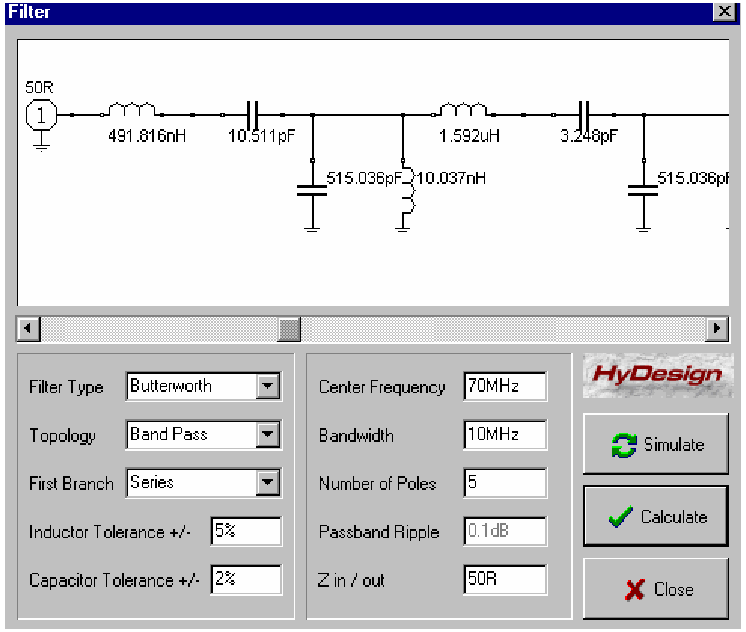
$$C_2 = \frac{1.618}{0.1432x50x438699539} = 515\mathbf{pF}$$

$$L_3 = \frac{2x50}{0.1432x438699539} = 1.59\boldsymbol{\mu}\mathbf{H}$$

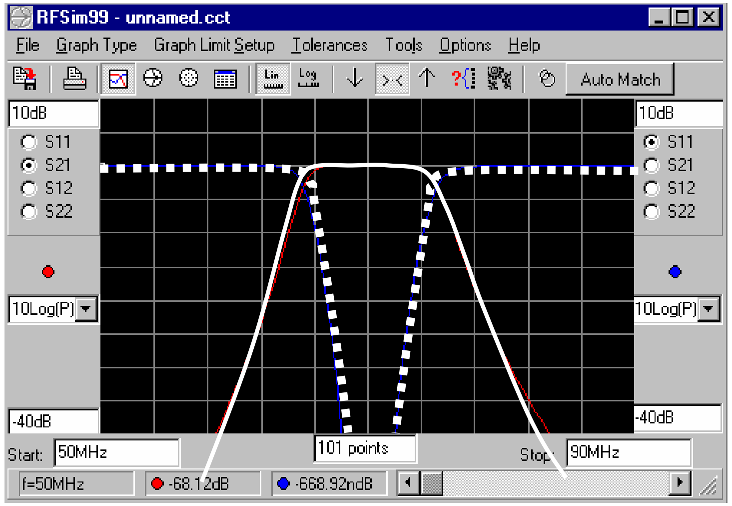
$$C_3 = \frac{0.1432}{2x50x438699539} = 3.3\mathbf{pF}$$

By symmetry, $C_5=C_1$, $L_5=L_1$ , $C_4=C_2$, $L_4=L_2$

This design is checked with “RFSim99” as shown in the screen dumps. Filter synthesis CAD programmes make the design much easier. Much more complicated filters are used in practice, in order to achieve objectives such as improved group delay flatness. Note that the values in this example have a large range; e.g. inductors from 10.1nH to 1.59uH. Such a large range is difficult to realise, and a different filter topology might be needed in practice……




**Synthesis Window**

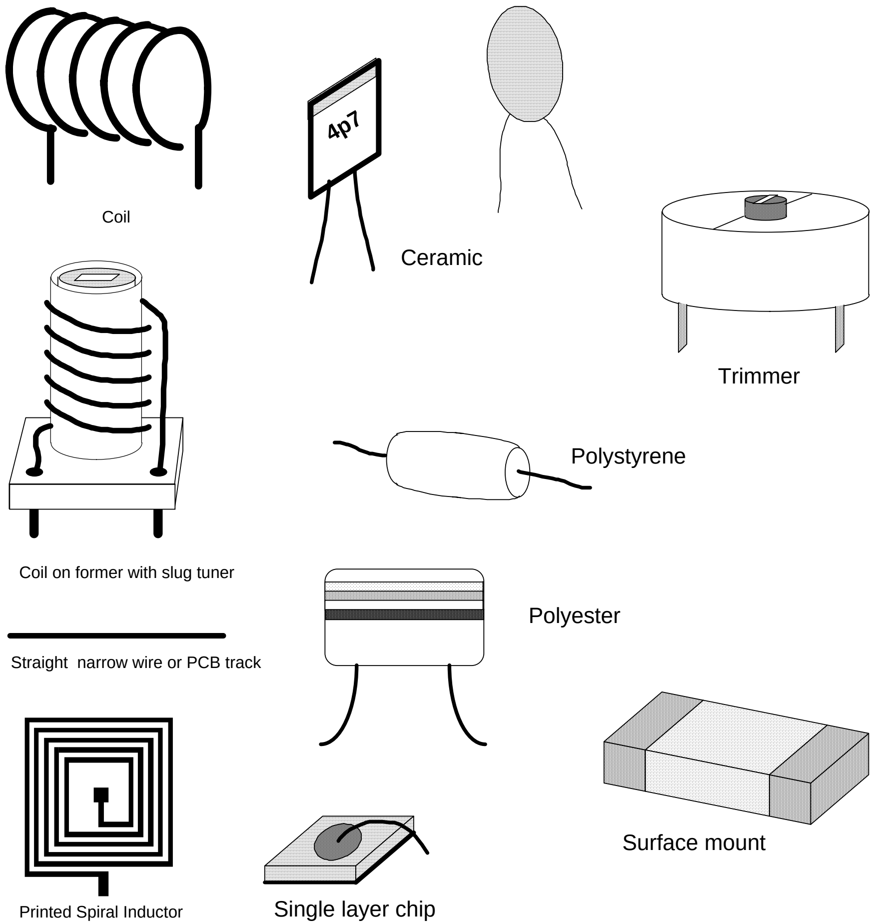



**Analysis Results**
**Lumped element filter synthesis with RFSim99 (available from RFGlobalnet)**



**PRACTICAL IMPLEMENTATION**

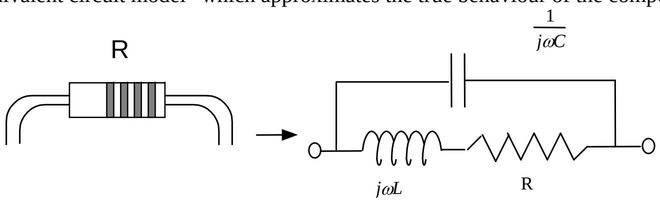
Having found the component values, the most suitable component types must be selected for the construction of the filter. The best component type depends on the design frequency and on the value of the component. The drawings show the sorts of component that should be selected for lumped-element RF filters.



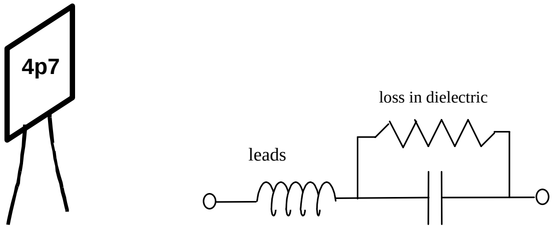

**Typical components for RF filters**

# PARASITICS

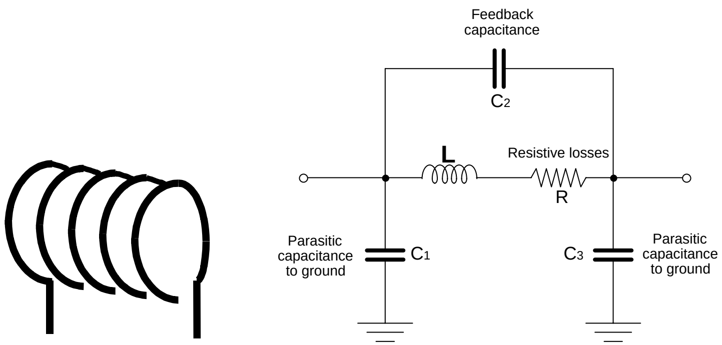
Real components have “parasitics” which include the inductance of leads, resistive losses in metals and dielectrics, fringing capacitance to ground, capacitance between turns, etc. These can be put into an “equivalent circuit model” which approximates the true behaviour of the component.




**Resistor model**




***Capacitor model***




***Inductor model***

The effects of these parasitics on your filter include increased insertion loss in the passband, reduced rejection in the stop band, reduced “sharpness” of transitions and spurious behaviour at high frequency (since every component has a self-resonant frequency, beyond which its behaviour is erratic). Ultimately, lumped elements become unusable at a certain frequency, and transmission-line circuits become the only way of realising a useful filter.



# TYPES OF HIGH FREQUENCY CIRCUIT

1. LOW FREQUENCY, HIGH PACKING DENSITY

Transistor gain is high so one can design OP-Amp type circuits (e.g. active filters)
Limited frequency range, as transistor capacitance causes roll-off

2. LUMPED ELEMENT

Use inductors and capacitors for filters or to “tune out” the transistor capacitances.
Limited frequency range due to parasitics of the lumped elements

3. TRANSMISSION LINE CIRCUITS

In RF transmission lines, signals are confined (**guided**) by CONDUCTORS. In these, “parasitics” are not a problem, but circuits are large below ~1 GHz for most applications.

# TRANSMISSION LINES

The purpose of a transmission line is to transfer a signal from a source over some distance to a remote load. In the case of electricity generation and distribution the signal is just a continuous sine wave and the aim is to get a great deal of power from the generator to the user as efficiently as possible. In computer networking, the signal is a pulse train of data. In radio systems the signal is one or more modulated carriers containing the information. The source might be a transmitter in a cabinet on the ground, and the load might be an antenna mounted at the top of a tower.


**General transmission line situation**

Note that the transmission line has a forward and a return path for current to flow. This is essential in order to obey Kirchoff’s laws.

The most important aspects of this transmission line are:-

1. The transmission line must be properly designed to have low loss and be shielded in some way to prevent the signal leaking away and to reject electrical interference from outside.
2. The line has some inductance and capacitance, distributed all the way along the line. This leads to the definition of a CHARACTERISTIC IMPEDANCE (more on this later).
3. There is a **significant time delay** from the signal leaving the source to it reaching the load.

These properties of the transmission line are summarised by the parameters of

**Characteristic Impedance**, Zo

**Propagation Constant,** $\gamma$. $\gamma$ has a real part $\alpha$ representing loss and an imaginary part $\beta$ representing the time delay. $\gamma = \alpha + j\beta$.

**Zo is a resistive impedance, but in the ideal lossless case there is no power dissipation associated with it. In circuit terms, as the signal flows down the line the power is temporarily stored in the magnetic field of the line inductance and the electric field of the line capacitance.**

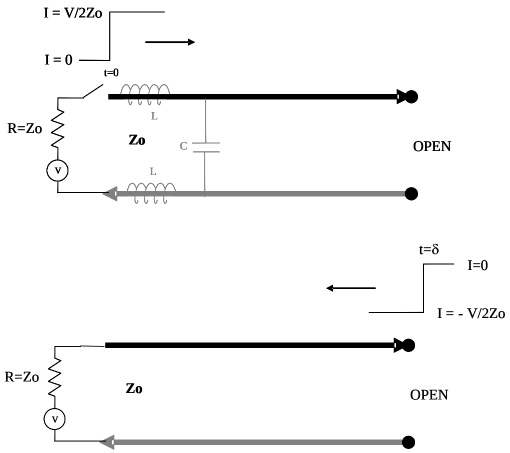
## REFLECTIONS ON TRANSMISSION LINES

As a result of the time taken for signals to travel down the transmission line, the source does not at first "know" what the impedance of the load is. If the voltage and current travelling down the line do not match the load impedance, a REFLECTION occurs at the load end. The reflection is created and is essential in order that Ohm's law is obeyed at the load end.

**Consider this example:** a voltage source V with source resistance R is connected by a switch to the transmission line of characteristic impedance Zo at time t=0. To get maximum power from the source into the TL, R is made equal to Zo. The load is an open circuit. Looking at the diagram, it should be obvious that the correct solution is that the current should be ZERO; but we only know that by cheating and looking immediately at the load end. The source can't do that, so initially a current starts to flow at t=0, with value V/2Zo (there is a potential divider effect between the source resistance and the Zo of the TL, giving ½ when R=Zo).

When this current step arrives at the load, the current has nowhere to go, so it is REFLECTED and a reverse step is created at time $t=\delta$, where $\delta$ is the time taken to travel down the line.

The value of the reverse step is –V/2Zo, and the two currents cancel out completely, satisfying the expected end result. So, there is some interesting TRANSIENT behaviour, but when everything settles down – in what is know as the "STEADY STATE" - we get the expected result of no current flowing.




**Example of reflections on a transmission line**

In ELECTRICTY POWER TRANSMISSION the transient behaviour can cause huge spikes and destroy equipment.
In COMPUTER NETWORKS the reflections cause data errors, as bits interfere with one another.
In RADIO SYSTEMS the reflections can also lead to damage to components, inefficient transfer of power and data corruption.

**<u>The way to avoid these problems is to ensure Zsource=Zload=Zo of the transmission line</u>**

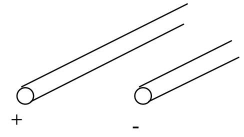
In RF & Microwave Engineering, 50 Ohms is the most commonly used impedance.

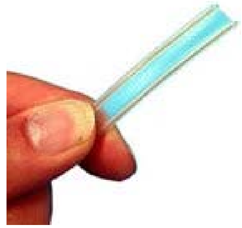


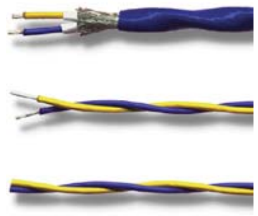
## PRACTICAL CONSTRUCTION OF TRANSMISSION LINES FOR RF

### Twisted Pairs and the Lecher Line

Twisted pairs started off life in telephony and were generally regarded as a cheap and simple means of achieving signal confinement for a low frequency transmission line. Nowadays they find widespread use in computer networking: "UTP" stands for Unshielded Twisted Pair and these cables are routinely used to 100Mb/s.



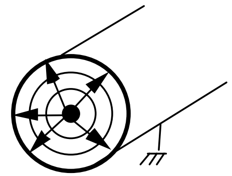


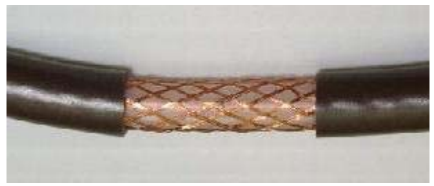
**300 Ohm FM ribbon from Waters and Stanton plc.**
http://www.wsplc.com

**Precision twisted pairs from Gore**
http://www.gore.com

### Coax

Coaxial lines consist of a centre conductor inside a cylindrical outer ground shield. Standard coax (like in the Teaching labs) is useable to a few hundred MHz. But higher precision types are useable up to 110 GHz.

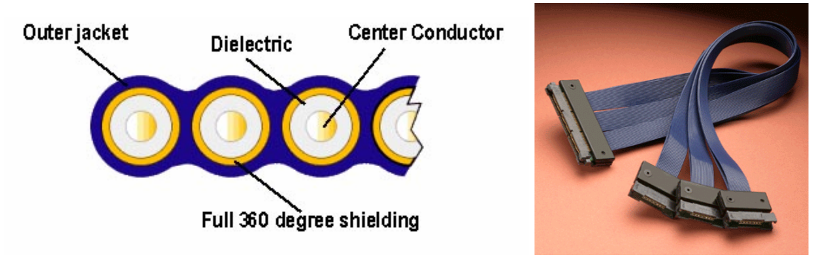




**Braided flexible coax**

**"SMA" connector**

As computers are operating beyond 3GHz, and high data rate connections are required without unwanted reflections, "Micro-coax" has started to make an appearance in ribbon cable form.




**"Blue Ribbon" Micro Coax from Tyco Electronics**
http://www.precisionint.com/HighSpeedData/BlueRibbon/

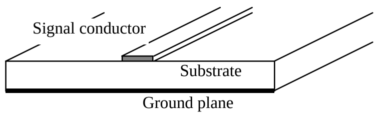
**Hollow Waveguide**

This is no longer a "two-wire" type of transmission line and we can't simply talk about forward and return current paths. The signal propagates as an electromagnetic wave, with a complicated field pattern. They are low loss and can handle high power.

**Some rectangular waveguide components with flanges (from www.Quinstar.com)**

**Microstrip**

This consists of a signal conductor mounted above a ground plane, usually by using a dielectric substrate. With special materials and precision fabrication, microstrip is useable to more than 100 GHz. On basic "PCB-like" materials it works to maybe a few GHz. It is fabricated with photomasks, so yields cheap components. However, losses are quite high giving moderate performance.

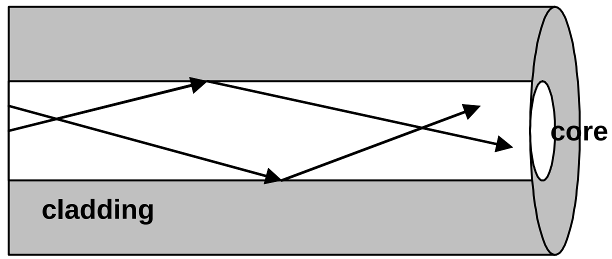



The Satellite TV LNB was an excellent example of a microstrip circuit.



**Optical Fibres**

These are more and more common in modern communications because of their enormous bandwidth and therefore information-carrying capacity. They are, of course, based on total internal reflection:-

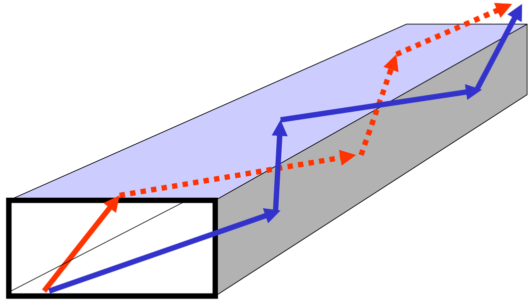



Note that multiple rays can travel down the fibre, and the fibre must ideally be designed in such a way that these rays all have the same propagation time down the fibre and they all arrive in-phase. Without this, data will be corrupted. Making sure the rays all superimpose properly is more properly referred to as designing the fibre to carry the correct *mode of propagation*. If the core is wide in diameter, rays can go off in all directions and many modes can exist. Most long-haul fibres have very narrow cores to ensure that only one mode can exist.

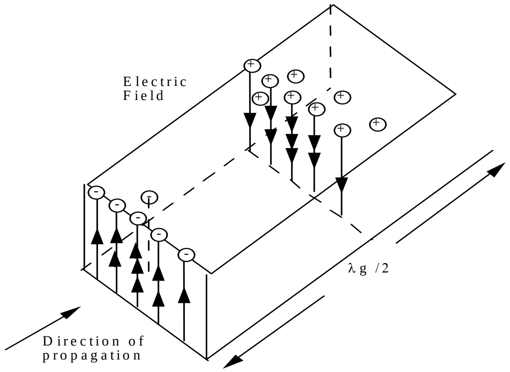
Fibres are key to the backbone of networks. They are also useful at this point for explaining the very difficulty subject of metallic microwave rectangular waveguides and their modes of propagation.

## RECTANGULAR WAVEGUIDES

These guide the signal by reflections off the metallic sidewalls. A rectangular waveguide is able to support a theoretically infinite number of modes. The one normally used is the TE10 mode, which is the one which starts to propagate at the lowest frequency. This can be visualised as the superposition of two em waves, bouncing back and forth down the waveguide at a critical angle such that they add constructively at the far end:-



The terms TE and TM stand for Transverse Electric and Transverse Magnetic. The mode indices (1,0 etc) in this case refer to the number of half cycles that fit across the waveguide in the *x* and *y* directions. z is the direction of propagation, along the longitudinal axis of the waveguide. In the TE10 mode, the E-field is always perpendicular to the direction of propagation, as illustrated below:-



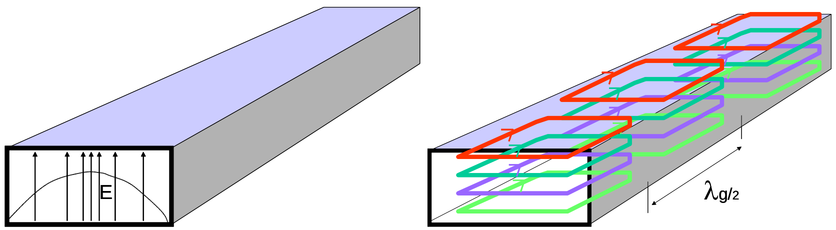




**Illustration of the E-field (left) and H-field (right) for the TE10 waveguide mode**



CUT-OFF FREQUENCY

Observing the E-field distribution from the previous page, it is necessary for a half-wavelength of the signal to be "fitted inside" the guide. This means that there is a LOWER CUT-OFF for the TE10 mode given by:-

$$a \le \frac{\lambda_{gc}}{2} \rightarrow f_c = \frac{V}{2a}$$

$$f_c = \frac{C}{2a\sqrt{\mu_r \varepsilon_r}}$$

Below this frequency, power cannot flow and the incoming wave is Evanescent

→ exponential decay with distance

→ only reactive power.

There is also an upper frequency limit because higher order modes can be excited; $TM_{11}$, $TE_{20}$, etc. Multiple modes = bad news. The mode velocities are different, so an information-carrying signal suffers massive pulse spreading (this is called DISPERSION).

Because of this multiple-mode characteristic, rectangular waveguides have a restricted useful bandwidth, and so a range of standard sizes is used to cover the whole microwave and millimetre-wave spectrum:-

**FREQUENCY BANDS**

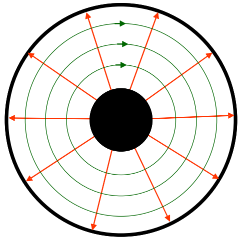
| | | | |
|---|---|---|---|
| L | 1 | to | 2 |
| S | 2 | to | 4 |
| C | 4 | to | 8 |
| X | 8 | to | 12 |
| Ku | 12 | to | 18 |
| K | 18 | to | 26.5 |
| Ka | 26.5 | to | 40 |
| Q | 33 | to | 50 |
| U | 40 | to | 60 |
| V | 50 | to | 75 |
| W | 75 | to | 110 |

> 1 GHz = microwave
> 30 GHz = mm waves or "EHF"
> 300 GHz = submm-waves…….and after that you move into infrared!

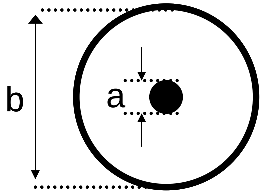
## COAXIAL LINES

Analysis of the electric and magnetic fields in the coaxial structure is reasonably straightforward because of its simple structure. The field is referred to as TEM (Transverse Electromagnetic) because the E and H field lines are both perpendicular to the direction of propagation. The H-field forms circular loops around the current in the centre conductor. The E-field forms radial lines between the centre conductor and the grounded outer conductor. The field **pattern** (not strength though) is exactly the same all the way down the cable.

Analysis yields the following characteristic impedance:-



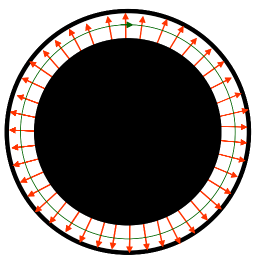
$$Z_o = \sqrt{\frac{L}{C}} = \frac{377}{2\pi\sqrt{\varepsilon_r}} \ln\left(\frac{b}{a}\right)$$

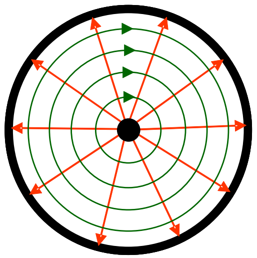



L and C are distributed along the cable. This is a vital concept in high frequency engineering that will be examined in detail later.

**COAXIAL 'CIRCUITS'**

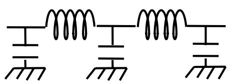
At low frequency, a high impedance line section will appear as a series inductance, whilst a low impedance section will act as a shunt capacitance:-

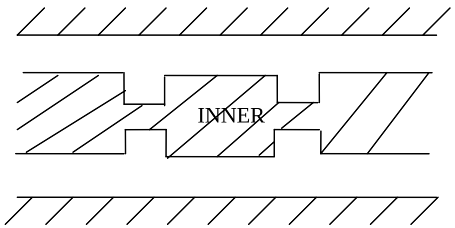

Low Z


High Z

So, a low pass filter can be realised with a machined centre conductor:-

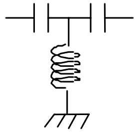
FILTERS:-

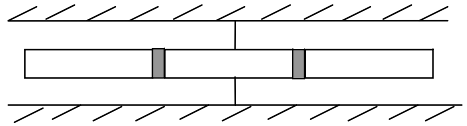




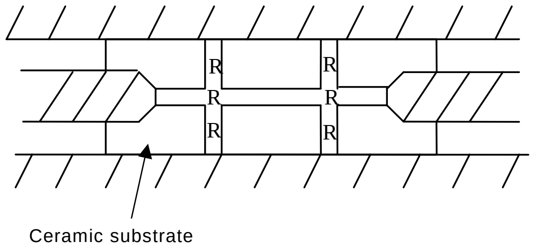



A **thin** disc shorting the centre conductor will act as a shunt inductor. A dielectric spacer in series in the centre conductor will act as a series capacitance. Hence, a high-pass filter can be realised:-





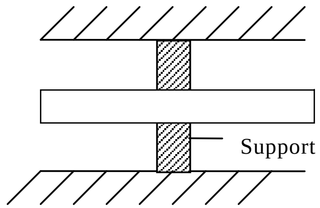
More complex components can be realised by inserting a ceramic substrate into the coax, with resistors printed on it, or even devices attached. The example shown below is an attenuator. Power sensors would use a similar structure, with a diode detector mounted on the substrate.




**Cross-section of a coaxial attenuator**

**<u>Frequency limitations of coaxial cables and connectors</u>**

Connectors with very narrow outer conductor diameter and air instead of dielectric give the highest frequency of operation. But even an air-based line must have supports:-




Poor ground continuity affects low frequency connectors like BNC ones, but even high quality microwave and millimetre-wave connectors and cables are limited in frequency range because of ***non TEM modes*** such as a transverse resonance in the support material, or transverse resonance in the air cavity. If the connector diameter is smaller, these resonances will be at higher frequencies. Hence, to achieve a higher maximum frequency, the inner diameter of the outer conductor must be made smaller. For example, to operate to 110GHz the 1 mm connector was developed in the mid 1990s.

**COMMON COAX CONNECTOR TYPES**

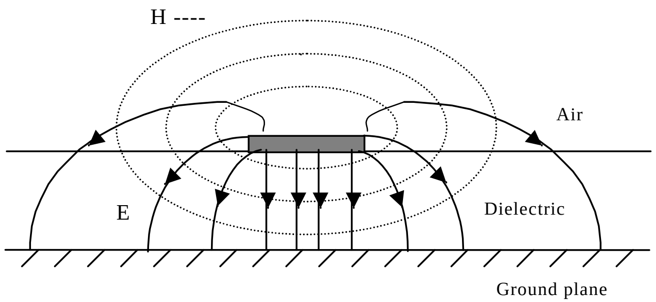
| | | | |
|---|---|---|---|
| BNC | < 1 GHz | APC7 | < 18 GHz |
| ‘N’ | < 18 GHz | SMA | < 18 GHz |
| APC2.4 | < 50 | APC3.5 | < 26.5 |
| ‘K’ | < 40 GHz | ‘V’ | < 65 |
| “W” , 1 mm | <110GHz | | |

APC3.5 is compatible with SMA and K
APC2.4 is compatible with V

MALE/FEMALE is determined by centre conductor. (In the US “plugs” and “jacks” are often referred too)

## MICROSTRIP

This is a simple low-cost technique for realising microwave transmission lines. The approximate field pattern is shown below. Since the dielectric is inhomogeneous, it does not support pure TEM fields, but quasi-TEM.




‘Microstrip’ is the most common for circuits because:-

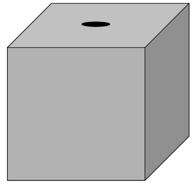
- Easy to make
- Can solder on components
- Can make filters, couplers and matching networks simply by ‘drawing’ them.



## TRANSMISSION-LINE RESONATORS

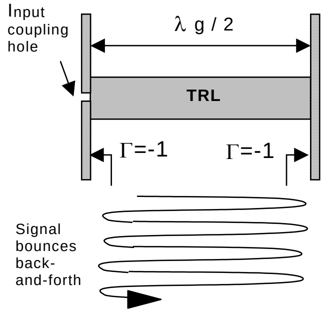
Broadly speaking transmission-line resonators are either based on reflection between a pair of open and/or short circuits, in a straight resonator, OR a circulating signal in a round resonator.

©ImagExtra



**Waveguide Cavity**

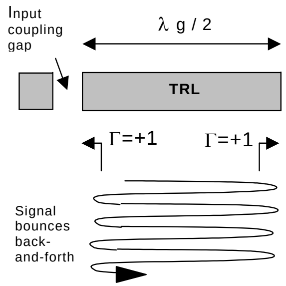
The principle is the same but the bottle resonance is from sound waves reflecting back-and-forth whereas the waveguide cavity resonance is due to electromagnetic waves reflecting back-and-for the from the metallic walls:




*e.g. waveguide cavity*

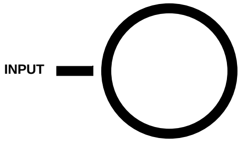
**HALF-WAVELENGTH RESONATOR WITH SHORT CIRCUIT ENDS**

This resonator has resonant frequencies given by $n\lambda_{g/2}$= physical length where $n$ is an integer




**HALF-WAVELENGTH RESONATOR WITH OPEN CIRCUIT ENDS**

This resonator has resonant frequencies given by $n\lambda_{g/2}$= physical length where $n$ is an integer.This is widely used in microstrip filters, since short-circuit resonators are hard to fabricate (holes need to be drilled in the substrate to connect to the ground plane).



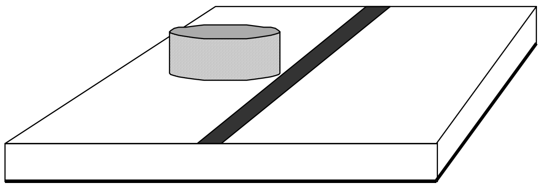

*e.g. microstrip ring*

**CIRCULAR RESONATOR**

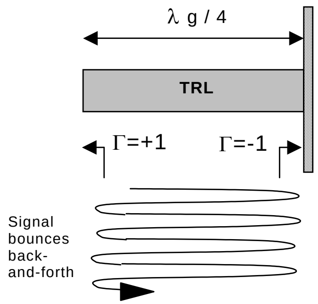
This resonator has resonant frequencies given by $n\lambda g = 2\pi$ x radius where $n$ is an integer.

i.e. the signal circulating round must travel an electrical distrance $n$ x 360 degrees each time around in order for it to add constructively.

Another important example is the DIELECTRIC PUCK; here, the reflection at each side is caused by total internal reflection at the boundary between the dielectric (which has high $\varepsilon r$) and air. The cylindrical puck actually has complex field patterns like a hollow waveguide.



**A dielectric puck, coupled to a microstrip line**




e.g. combline filter

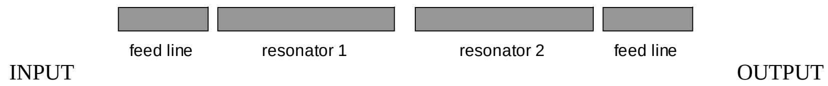
**QUARTER-WAVELENGTH RESONATOR; one end open, the other short circuit**

This resonator has resonant frequencies given by $n\lambda g_{/4}$= physical length where $n$ is an integer



In all cases, the signal is coupled in and out through something like a hole in a ground plane, or a gap between signal conductors. It is important that the LOADED Q is not lowered too much by overcoupling. Ultimately, the loss in the transmission line determines the highest Q-factor that can be achieved.

A complete bandpass filter can be constructed by coupling a number of resonators together; e.g.:-

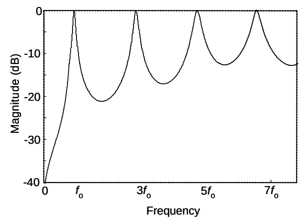



**Bandpass Filter using End-Coupled Half-Wavelength Open-Circuited Resonators**

Note that this consists only of capacitively-coupled resonators, and cannot be designed directly with the simple synthesis method shown earlier for LC filters.

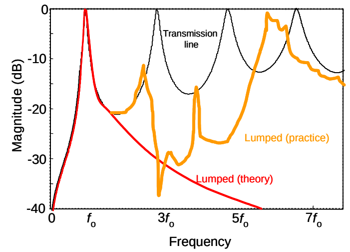
**Filter Example: A 10 GHz bandpass filter on alumina (12 x 12 mm)**

The resonators are half-wavelength long and open at each end.
They are coupled together side by side.




Frequency response of open/short quarter-wave resonator

Swindled? The whole point of replacing lumped elements was to get proper high frequency behaviour. The key difference of distributed elements is that this harmonic behaviour is extremely predictable, whereas lumped element behaviour is almost random at high frequencies:-



## S-PARAMETER EXAMPLE

A filter is a useful example to give an insight into S-parameter values and the wave behaviour of circuits. First, the deciBel units need to be understood.

**dB dB dB dB dB dB dB dB dB dB dB dB**

- dB is a unit based on the logarithmic ratio of powers. dB units are able to conveniently represent a huge range of signal levels and high gain/attenuation ratio: e.g. from pW to 100's of Watts is often encountered in radio.

$$\text{Gain in dB} = 10 \log \frac{\text{OUTPUT POWER}}{\text{INPUT POWER}}$$ , for an amplifier for example

Note that this is a ratio and has no absolute units.

Since power = $V^2/R$, it is important to remember to use 20, not 10 log for voltages:-

$$\text{Gain in dB} = 20 \log \frac{\text{OUTPUT VOLTAGE}}{\text{INPUT VOLTAGE}}$$ p-p, rms, or amplitude; it's a ratio

### S-parameters in decibels

Because S-parameters are voltage-based, to get dB, you must use:
**20 log (magnitude of Snm)**.

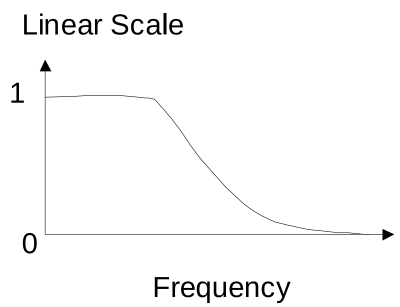
For example, the power gain in dB of an amplifier is **20 log (mag S21)**
The return loss is a figure of matching quality and is **20 log (mag S11)**

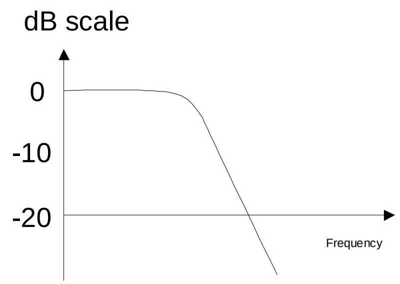
Hence, for the matched case S11=0, which gives S11=-infinity dB
For the open or short circuit, mag (S11)=1, so mag( S11)=0 dB (only the phase is different, and in both the short and open circuit case all of the signal is reflected)

**The input or output power must have some absolute units:-**

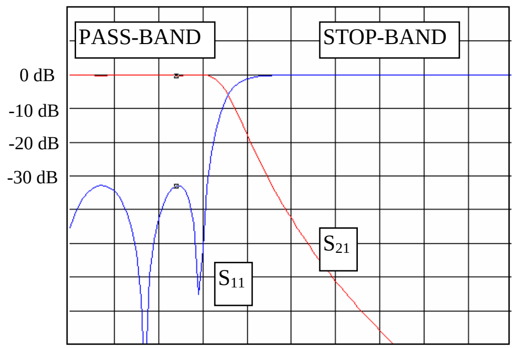
$$Power\ in\ dBW = 10\log(P\ in\ Watts)$$
$$Power\ in\ dBm = 10\log(P\ in\ mW)$$

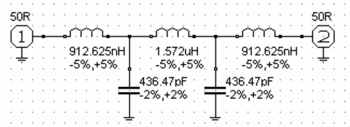
Compare a LPF in linear units and then in dB:-



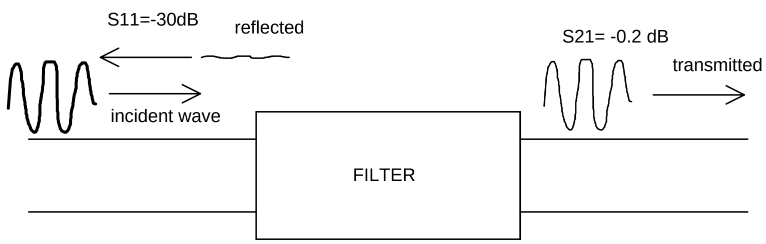



*Chebychev Low-Pass Filter Example*

***IN THE PASS-BAND THIS IS HAPPENING:-***




IN THE STOP-BAND THIS IS HAPPENING:-