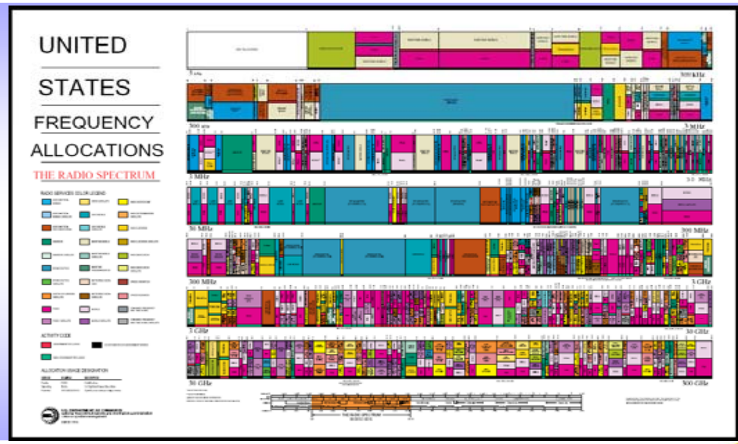
Lecture #1

Passives

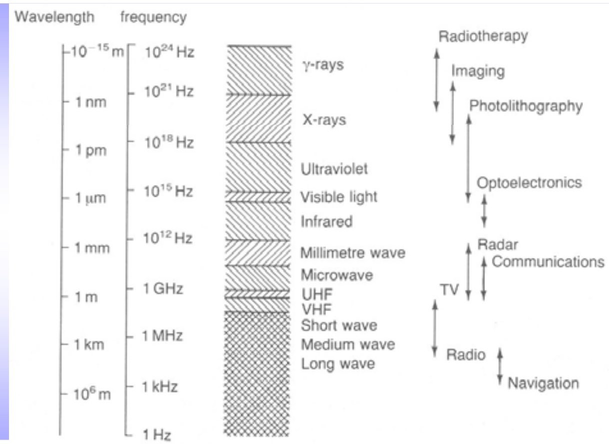
The main objective of this lecture is to introduce passive components, discuss their technological limitations and give modern examples of how they can be realised. The basic concepts outlined here will also act as a basic introduction to subsequent lectures.

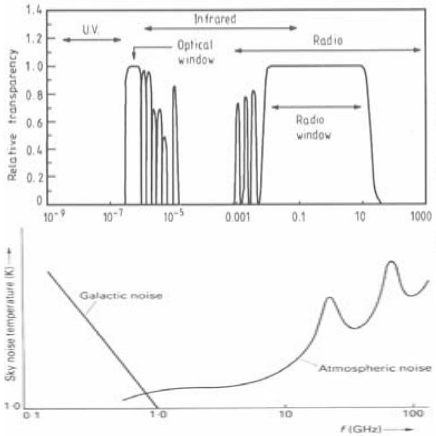
Frequency Spectrum and Size

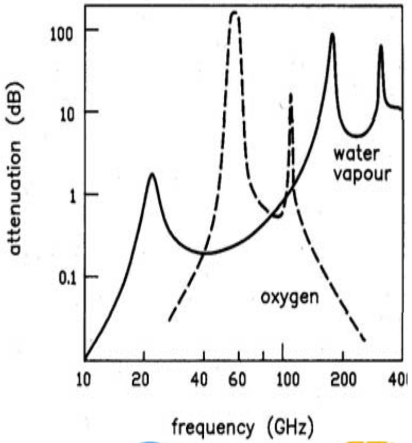


OVERVIEW

- Frequency spectrum and size
- Conventional RF materials
- Classical skin depth
- Distributed-element components
- Directional couplers
- Basic transmission line theory
- Transmission-line resonators
- Radiating structures

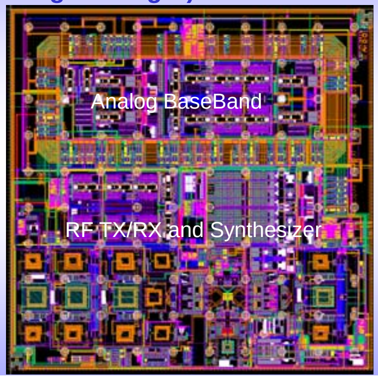




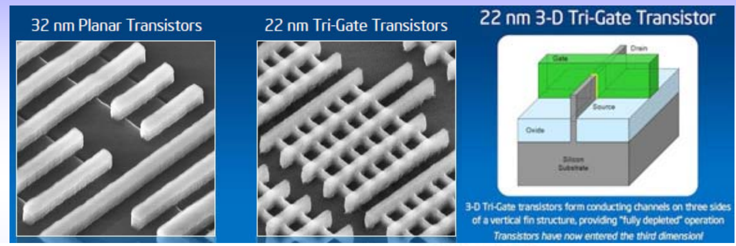


Intel's Cloud Computer - 48 Pentiums on a Chip using 45 nm node technology

2000 Intel RF CMOS Transceiver (70 nm gate length) for the 5 GHz Band



Intel's brand new 22 nm node technology



Process Name	P1266	P1268	P1270	P1272	P1274
Lithography	45 nm	32 nm	22 nm	14 nm	10 nm
1st Production	2007	2009	2011	2013	2015

Conventional RF Materials

Q. Why Bother to Understand Materials?

A. Commercial analogue integrated circuits are now operating up to 76.5 GHz . With CMOS compatible on-chip optical interconnect, the clock frequency at the 22 nm node technology is assumed to be 36.3GHz (Chen *et al.*, 2007), with their 3rd, 5th and 7th harmonics at 109, 182 and 254 GHz, respectively. Therefore, it is important to know the material characteristics from DC up to these frequencies, and beyond, so that the performance of these circuits can be fully characterised:

Permeability,	$\mu = \mu_0\mu_r$ [H/m]
where,	$\mu_0 = 4\pi 10^{-7}$ [H/m]
	μ_r = relative permeability ($\mu_r = 1$ in free space and ≈ 1 in non-magnetic materials, e.g. most dielectrics and good conductors)
Permittivity,	$\varepsilon = \varepsilon_0\varepsilon_r$ [F/m]
	$\varepsilon \equiv \varepsilon' - j\,\varepsilon''$
	$\varepsilon_0 \approx 8.854 \times 10^{-12}$ [F/m]
	ε_r = relative permittivity
and,	$\varepsilon_r \equiv \varepsilon_r' - j\,\varepsilon_r''$
where,	ε_r' = dielectric constant (ε_r' is quoted by the manufacturer for dielectrics, and $\varepsilon_r' = 1$ for free space)
Conductivity,	$\sigma \equiv \sigma' - j\,\sigma''$ [S/m] (σ' is quoted by the manufacturer for metals at DC, σ_o)
Effective Conductivity,	$\sigma_{eff} = \sigma + j\,\omega\varepsilon \equiv j\omega\varepsilon_{eff}$ [S/m]
Effective Permittivity	$\varepsilon_{eff} = \varepsilon - j\,\sigma/\omega$
therefore,	$\sigma_{eff}' = \omega\varepsilon_0\varepsilon_{r\,eff}''$
	$\varepsilon_{r\,eff}' = -\sigma_{eff}''/\omega\varepsilon_0$ (for a metal, this is sometimes known as the dielectric function and is a negative number)
and,	$\varepsilon_{r\,eff}'' = \sigma_{eff}'/\omega\varepsilon_0$
Loss Tangent,	$\tan\delta = \dfrac{\varepsilon_{r\,eff}''}{\lvert\varepsilon_{r\,eff}'\rvert}$ (this value is quoted by the manufacturer for dielectrics)

$$\tan\delta = \left| \begin{array}{l} 1\times10^{-4} \textit{ with alumina@10GHz} \\ 5\times10^{-2} \textit{ with polyimide@10GHz} \end{array} \right. \qquad \text{Dielectric Quality Factor, } Q = \frac{1}{\tan\delta}$$

Velocities Associated with Conductors

1.With a transmission line supporting TEM-mode propagation, the E- & H-field waves propagate outside the conductors at the phase velocity inside the dielectric, v_{pd}:

$$Vpd = \frac{c}{\sqrt{\mu_r\varepsilon_r}} \rightarrow c \cong 3.0x10^8 \ m/s \quad \textit{in vacuum}$$

where, c = speed of light in a vacuum; μ_r = relative permeability of the surrounding dielectric; ε_r = relative permittivity of the surrounding dielectric

2. The free electrons inside the conductors travel at the Fermi speed, v_f:

$$v_f = \sqrt{\frac{2E_f}{m}} \cong 1.6x10^6 \ m/s \quad \textit{for copper}$$

where, E_f = Fermi Energy; m = mass of the electron

3. The E- & H-field waves propagate inside the conductors at the phase velocity, v_{pc}:

$$v_{pc} = \frac{\omega}{\beta} = \omega\delta = \sqrt{\frac{2\omega}{\mu_o\sigma_o}} \cong 4.1x10^4 \ m/s \quad \textit{for copper at } 10 \textit{ GHz}$$

4. Conduction current, $Jc = N\,e\,v_d$, flows when (N e) electrons drift at a time average drift velocity, v_d:

$$v_d = \frac{I/Area}{N \cdot e} \cong 1.8x10^{-4} \ m/s \quad \textit{with } 10A \textit{ flowing in } 2.25 \textit{ mm dimeter copper wire}$$

A plane wave is defined as an electromagnetic wave having a wavefront with space quadrature E- & H- fields that are mutually orthogonal with the direction of propagation. Within a simple guided-wave structure, a plane wave is called a pure transverse electromagnetic (TEM) wave. A TEM wave propagating in one-dimension space can be completely described by the E-field component:

$$E(t,z) = E(0,0)e^{-\alpha z}e^{j(\omega t - \beta z)} \qquad E(t,z) = E(0,0)e^{j\omega t}e^{-\gamma z} \quad [V/m] \qquad H(t,z) = \frac{E(t,z)}{\eta} \quad [A/m]$$

Propagation Constant, $\gamma \equiv \alpha + j\beta$

where,	α = attenuation constant [Np/m]
	β = phase constant [rad/m]
also,	$\gamma = j\,$km
where,	modified wavenumber, km $= 2\pi$k,
where,	k = wavenumber (note that Re{k} gives the number of wavelengths per meter)

$$k = \frac{1}{\lambda} \qquad \textit{where, wavelength, } \lambda = \frac{\textit{phase velocity, } v_p}{\textit{frequency, } f} \quad [\mathrm{m}]$$

$$\gamma = j\frac{\omega}{v_p} \quad \textit{where, } v_p = \frac{1}{\sqrt{\mu\varepsilon}} \qquad v_p = \frac{c}{\sqrt{\varepsilon_r'}} \qquad \gamma = j\beta_o\sqrt{\varepsilon_r'(1 - j\tan\delta)}$$

Now, the magnetic field strength, H(t,z), is related to the electric field strength by the intrinsic impedance of the material, η:

$$H(t,z) = \frac{E(t,z)}{\eta} \quad [A/m]$$

Note that this is the electromagnetic representation of Ohms law!!!!!!!

Intrinsic Impedance

$$\eta = \sqrt{\frac{\mu}{\varepsilon}} \quad [\Omega] \qquad \left(c.f. \textit{ Characteristic Impedance of a lossless transmission line, } Zo = \sqrt{\frac{L}{C}} \quad [\Omega] \right)$$

now, $\eta_o = \sqrt{\dfrac{\mu_o}{\varepsilon_o}} \approx 120\pi \quad [\Omega]$ $\quad \eta_0$ = intrinsic impedance of free space

$$\eta = \frac{\eta_o}{\sqrt{\varepsilon_r'(1 - j\tan\delta)}}$$

Note that, $\quad \gamma = \dfrac{j\omega\mu}{\eta}$

Therefore, for any material, one only needs to know either γ or η and the other can then be calculated.

Low Loss Conductors ($\therefore \tan\delta >> 1$)

$$\gamma \cong \sqrt{\frac{\omega\mu\sigma'}{2}}(1+j)$$

since, $\tan\delta >> 1$

$$\alpha \cong \beta$$

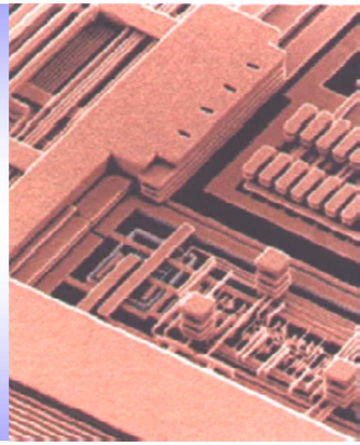
$$\gamma \equiv \frac{j\omega\mu}{\eta}$$

$$Zs_o = \sqrt{\frac{\omega\mu}{2\sigma_o}}(1+j) \equiv \eta$$

Lossless Dielectric, Lossy Conductor

Surface Power Density, $P_S = |J_S|^2\ Rs \quad [\mathrm{W/m^2}]$

where, J_S = Surface Current Density [A/m]



IBM's CMOS 7S ASIC technology

Metals of Interest

Only very high electrical conductivity metals are of interest for carrying signals, using low loss RF interconnects and transmission line structures. Only the following will be considered:

(1) **Silver:** coined silver (90% Ag, 10% Cu) is used to line the inside walls of ultra-low-loss mm-wave MPRWGs

(2) **Copper:** used to line the inside walls of microwave MPRWGs and is also the main metal used in hybrid microwave ICs and IBM's CMOS 7S ASIC technology

(3) **Gold:** used to line the inside walls of low-loss mm-wave MPRWGs, the main metal used in hybrid and monolithic (GaAs & InP) microwave/mm-wave ICs, bond-wire/strap interconnects, electrical contact coatings for low-loss mm-wave connectors and reflective coatings for optical mirrors & switches at infrared frequencies

(4) **Aluminium:** the main metal used in monolithic (silicon & SiGe) microwave/mm-wave ICs and has the highest conductivity of any metal at the boiling point of liquid nitrogen (i.e. 77 K).

Low Loss Dielectrics ($\therefore \tan\delta << 1$)

$$\gamma \cong 2\pi k(0.5\tan\delta + j) \quad \textit{since, } \tan\delta << 1$$

$$\beta \cong \omega\sqrt{\mu\varepsilon'} = km_o\sqrt{\varepsilon_r'} = km$$

$$\left\{ \textit{Note that, } \beta \cong km\left[1+\frac{(\tan\delta)^2}{8}\right] \quad \textit{when Bionomial Series is further expanded} \right\}$$

$$\textit{also, } \eta = \sqrt{\frac{\mu/\varepsilon'}{1-j\tan\delta}} \cong \frac{\eta_0}{\sqrt{\varepsilon_r'}(1-j\tan\delta/2)} \quad \textit{since, } \tan\delta << 1$$

Lossy Dielectric, Lossless Conductor

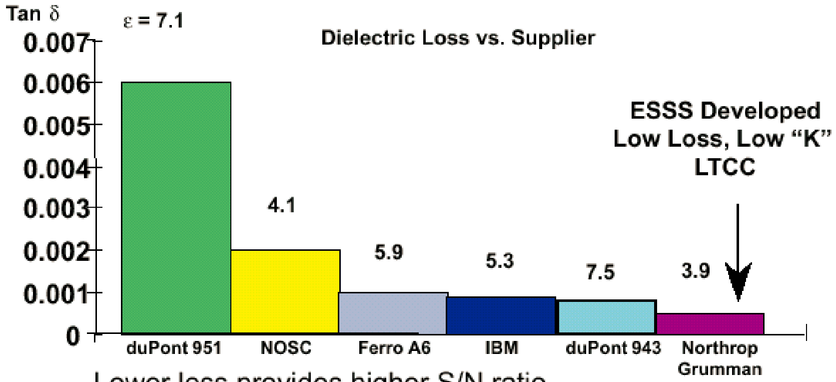
$$Q=\frac{\beta}{2\alpha},\quad where\quad \beta=\frac{2\pi}{\lambda g}\quad \therefore Q=\frac{\pi}{\alpha\lambda g}$$

$$where,\ attenuation\ per\ unit\ guided\text{–}wavelength=\alpha\lambda g\quad [Np/\lambda g]$$

$$\{Note\ that,\ Power\ Attenuation=10\log_{10}\left(e^{-2\alpha\lambda g}\right)=-20\alpha\lambda g\log_{10}(e)=-8.686\ \alpha\lambda g\quad [dB/\lambda g]\}$$

$$For\ a\ low\ loss\ dispersionless\ dielectric:\quad \alpha=\frac{\sigma'}{2}\sqrt{\frac{\mu}{\varepsilon'}}=f(\omega)\ \ since\ \ \sigma'=\omega\varepsilon''\quad and\quad \beta=\omega\sqrt{\mu\varepsilon'}=f(\omega)$$

$$therefore,\ \alpha\lambda g=\frac{\pi\sigma'}{\omega\varepsilon'}\neq f(\omega)\quad and\ tan\delta=\frac{\varepsilon''}{\varepsilon'}\neq f(\omega)\quad \therefore Q=\frac{\omega\varepsilon'}{\sigma'}\equiv\frac{1}{tan\delta}\neq f(\omega)$$



Lower loss provides higher S/N ratio
Lower dielectric constant enables easier manufacturability

e.g. For polyimide, having a relative permittivity $\varepsilon_r = 2.8 - j\,0.039$, the following can be calculated for a frequency of 30 GHz:-

$$\tan\delta=\frac{\varepsilon r''}{\varepsilon r'}=0.014\qquad \lambda=\frac{c}{f\sqrt{\varepsilon r'}}=6\,[mm]$$

$$Q=\frac{1}{\tan\delta}=72$$

$$k=\frac{1}{\lambda}=167\,[\lambda/m]$$

$$\gamma=1051(0.5\tan\delta+j)=7.36\,[Np/m]+j1051\,[rad/m]$$

$$Power\ \ Attenuation=e^{-2(\alpha l)}$$

$$\alpha\lambda=0.044[Np/\lambda]$$

$$Power\ \ Attenuation=\alpha\lambda(20\log_{10}e)=8.686\alpha\lambda=0.3836\,[dB/\lambda]$$

$$\eta=\frac{j\omega\mu}{\gamma}\quad were\ \omega=2\pi f[rad/s]\quad and\ \ \mu=\mu o=4\pi10^{-7}[H/m]$$

$$\eta=225+j1.58\Omega$$

Classical Skin Depth

DC Conduction Current in a Wire

The net charge at any point within the wire is always zero. This is because as soon as a negatively charged free electron detaches from its normally neutrally charged donor atom the remaining positively charged fixed ion quickly attracts a free electron from a neighbouring atom. As a result, a ‘sea’ of free electrons is said to exist in the conductor.

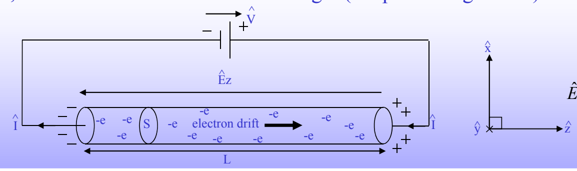
Conduction Current, I = Jc S [A]

where, Jc = conduction current density [A/m^2]

and, S = cross-sectional area [m^2]

Now, Jc = σ_o Ez (where, σ_o = Neμ and electron mobility, μ = τe/m)

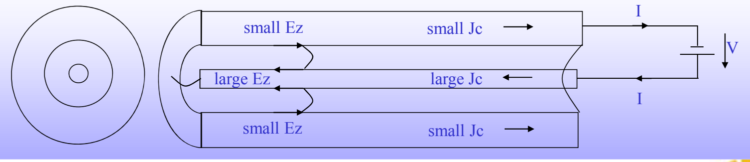
where, Ez = electric field strength (i.e. potential gradient) in the z-direction



$$\hat{E}z=-\frac{\partial\hat{V}}{\partial z}\equiv-\frac{\hat{V}}{L}\ at\ DC$$

DC Conduction Current in a Coaxial Cable

The conduction currents flowing in the inner and outer conductors are identical. However, the conduction current densities are not the same, since the cross-sectional areas are different. If the two conductors are at a different potential there will be equal but opposite surface charges at the two conducting boundaries, which will induce an E-field component normal to the conductor surfaces ($En = xEx + yEy$). The tangential electric field strengths inside the two conductors are going to be different because they are directly proportional to the conduction current densities. Therefore, since Ez is larger inside the inner conductor, the E-field will tilt more forward at the inner conductor surface than at the outer conductor surface.



Boundary Conditions with $\sigma_o \rightarrow \infty$:

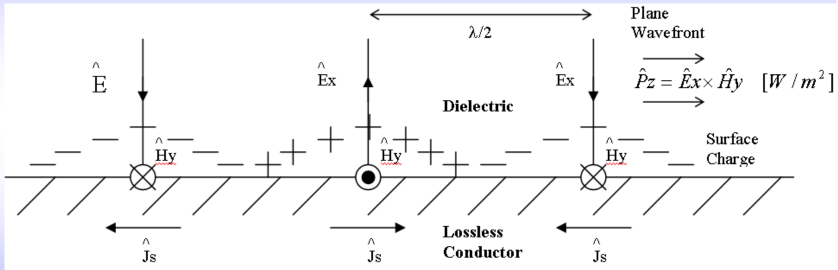
1/ There are no E-fields within the conductor, since there are no gradients of potential. Therefore, the E-field which is tangential to the conducting plane ($Et = yEy + zEz$) must vanish at the boundary. Moreover, no power is dissipated in a perfectly conducting plane, since there are no E-fields within the conductor and, therefore, electromagnetic waves cannot exist.

2/ Any external E-field normal to the plane ($En = xEx$) must be terminated by a surface charge having a surface charge density, $Qs = n \cdot Ds$ [C/m^2], where surface charge displacement, $Ds = \varepsilon En$ [C/m^2]. Therefore, the spatial distribution of Qs corresponds to that of Ex.

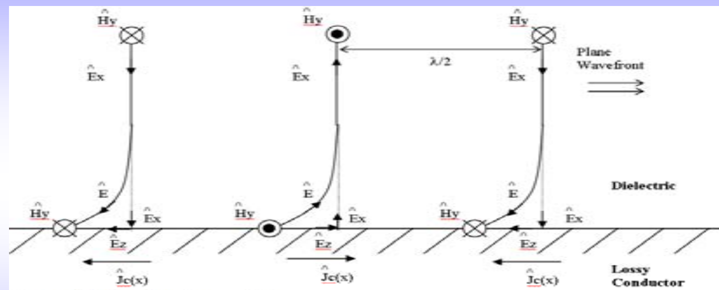
3/ An H-field that is tangential to the conducting plane ($Ht = yHy + zHz$) will induce a surface current with a density that is equal to the magnetic field strength, $Js = n \times Ht$. Therefore, the spatial distribution of Js also corresponds to that of Ex, since $Hy = Ex/\eta$.

4/ An H-field that is normal to the conducting plane ($Hn = xHx$) must vanish at the boundary.

RF Conduction Currents in a Lossless Conducting Plane



RF Conduction Currents in a Lossy Conducting Plane



Leontovich direct boundary condition:

$$\bar{E} \times \hat{n} = Zs\, \hat{n} \times (\bar{H} \times \hat{n})$$

where $\hat{n}$ denotes the outward local surface unit vector and Zs is the Leontovich Surface Impedance

Now, $Ez(x{=}0) = Zs_o\, Js$

Therefore, when σ_o is made finite, a tangential E-field exists, since the surface impedance is no longer zero. Also, since $Ez(0)$ exists, the resultant E-field in the dielectric leans forward, just above the surface of the conductor, i.e. $E = xEx + zEz$. Therefore, just above the surface, the wave is not pure TEM, because the E-field, H-field and direction of propagation are not mutually orthogonal. Now, since Ez and Hy exists inside the conductor, a wave can propagate inside this material, i.e. with Poynting vector $Px(x) = Ez(x) \times Hy(x)$, where $Ez(x) = Ez(0)e^{-\gamma x}$ and $Hy = Ez/Zs_o$.

If a wave propagates inside the metal, the associated E-field will induce a conduction current, $Jc(x) = \sigma_o Ez(x) = Jc(0)e^{-\gamma x}$. At the surface of the conductor, the conduction current leads the surface current by 45°, since $Jc(0) = \sigma_o Zso\, Js = \sqrt{2}\, \sigma_o Rso\, Js\, e^{+j\pi/4}$.

Now, it can be shown that: $Js = \int_0^\infty Jc(x).dx$

Js is only a theoretical concept but, in practice, its value does not vary much as σ_o reduces in value. If the time dependency is ignored:

$$\hat{J}s = \frac{\hat{J}c(0)}{-\gamma}\left[e^{-\gamma x}\right]_0^\infty \qquad \gamma = \sigma_o Zs_o$$

$$\therefore \hat{J}c(0) = \gamma \hat{J}s$$

The pointing vector $Px(x) = Ez(x) \times Hy(x)$, where $Ez(x) = Ez(0)e^{-\gamma x}$ and $Hy = Ez/Zs$. Therefore, the time-average power dissipated at any depth inside the metal is represented by the power flux density, $P_D(x) = Re\{Ez(x)\, Hy^*(x)\}$.

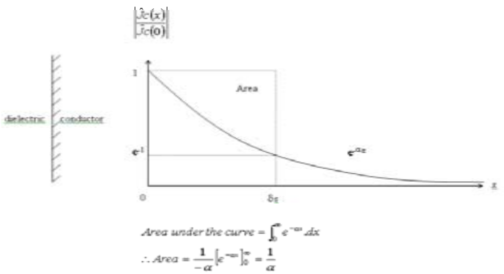
As a result, power density, normalised to its surface value, is equal to $e^{-2\alpha x}$.

At one skin depth, the power density has decreased by 8.686 dB from its surface value.

In practice, when a metal sheet is used to screen electromagnetic radiation, the recommended thickness is between 3 and 5 skin depths.

Skin Depth in Metal Conductors

It has been shown that the magnitudes of the E-field and H-field components of an electromagnetic wave, and the conduction current density inside a conductor, all decay exponentially with distance into a material, i.e. with $e^{-\alpha x}$.



$$\delta s = \sqrt{\frac{2}{\omega \mu_o \sigma_o}} \equiv \delta s_o$$

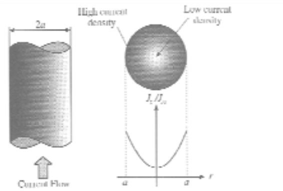
$$Rs_o = \frac{1}{\sigma_o \delta s_o}$$

$$also,\ \delta s = \frac{(1+j)}{\gamma}$$

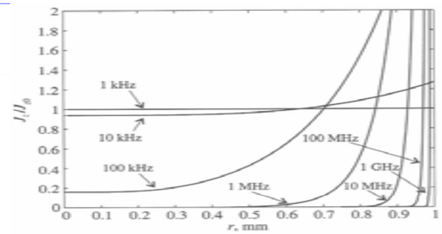
$$\therefore \hat{J}c(0) = \frac{\hat{J}s}{\delta s}(1+j)$$

Within a good conductor, $\lambda = 2\pi\delta s$

At one skin depth, power density decreases by 8.686 dB from its surface value. When used to screen electromagnetic radiation, the recommended thickness is between 3 and 5 skin depths.

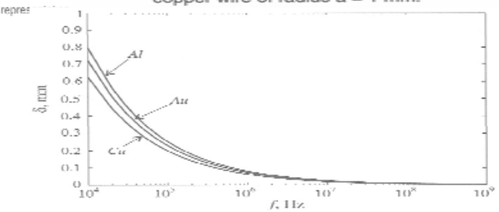


Schematic cross-sectional AC current density repres... normalized to DC current density.

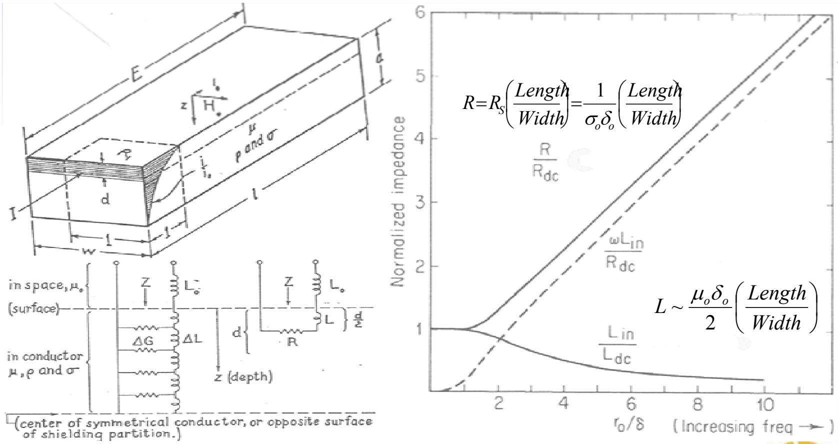


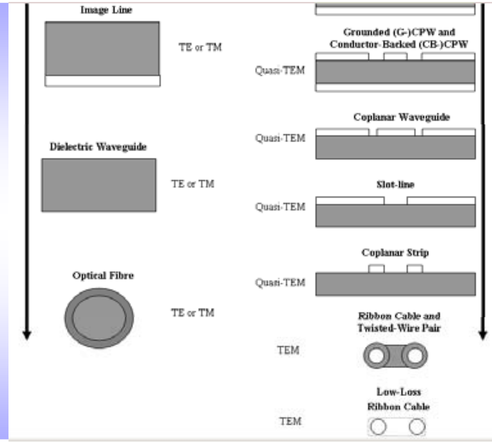
b) Frequency behavior of normalized AC current density for a copper wire of radius $a = 1$ mm.

...re 1-6 Increase in resistance over the cross sectional surface area. The current flow is confined to a small area defined by the skin depth δ.

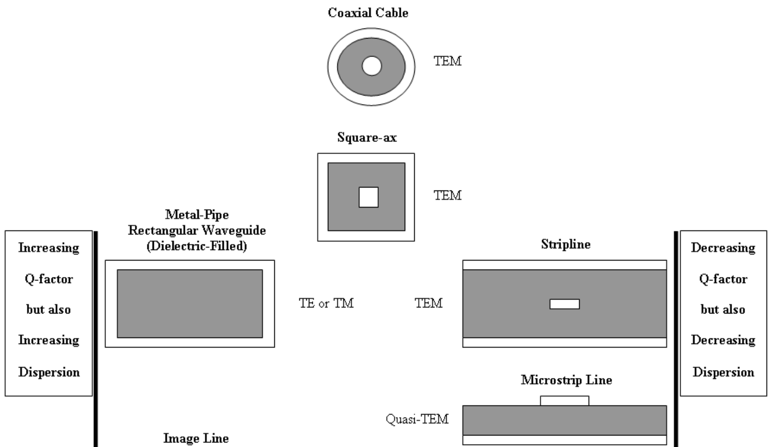


Skin depth behavior of copper $\sigma_{Cu} = 64.516\times10^6$ S/m, aluminum $\sigma_{Al} = 40.0\times10^6$ S/m, and gold $\sigma_{Au} = 48.544\times10^6$ S/m.





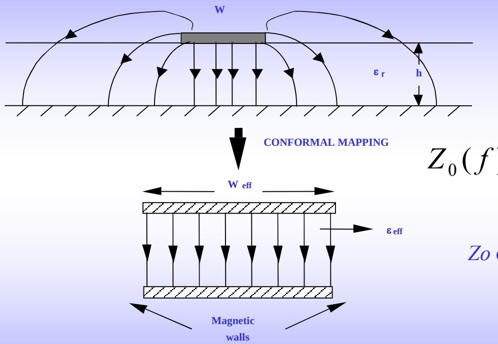
Distributed-element Components



Distributed-element Techniques

- Beyond 20 GHz the spiral inductors are beyond their useful frequency range, because of their own self-resonance, and so distributed elements are used for matching
- Distributed elements can be realised in a number of transmission-line media, with microstrip and CPW being by far the most common
- Highest operating frequency of an MMIC is limited only by the maximum frequency at which the active devices still have usable available gain
- Lowest frequency of operation is determined by the chip size, since the physical length of matching elements is too great at frequencies below ~ 5 GHz

The Quasi-TEM 'Magnetic Wall' Model



$$Z_0(f) = \frac{h\eta}{W_{eff}(f)\sqrt{\varepsilon_{eff}(f)}}$$

Zo changes slightly with frequency

By reverse-mapping these field lines into the original microstrip domain, the microstrip fields are found

Microstrip Frequency Limitations

1. The lowest-order transverse microstrip resonant mode

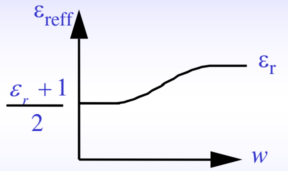
* occurs when the width of the line (plus a fringing field component) approaches a half-wavelength in the dielectric:-

$$f_{CT} = \frac{c}{2\sqrt{\varepsilon_r}(w+2d)} \quad where \quad d = 0.2h$$

$$f_{CT} \cong 52\ GHz\ \ with\ 50\ \Omega\ lines\ on\ 635\ \mu m\ \ thick\ alumina$$

Avoid wide lines or introduce slots!!!

ε_{eff} is found to vary width strip width



$$\varepsilon_{r\,eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{h}{w}\right)^{-0.555}$$

* Narrow lines, field is equally in substrate and air
* Wide lines, it is similar to a parallel-plate capacitor
* ε_{eff} also changes with frequency because the field profile changes

2. TM mode propagation

* A frequency for strong coupling between the quasi-TEM mode and the lowest-order TM mode occurs when their phase velocities are close to one another:

$$f_{C(TEM,TM)} = \begin{cases} \dfrac{c\tan^{-1}(\varepsilon_r)}{\sqrt{2}\pi h\sqrt{\varepsilon_r - 1}} \cong 50\ GHz\ \ with\ 50\ \Omega\ lines\ on\ 635\ \mu m\ thick\ alumina \\ \approx \dfrac{c}{4h\sqrt{\varepsilon_r - 1}} = \dfrac{75\ [GHz]}{h[mm]\sqrt{\varepsilon_r - 1}} \quad for\ wide\ lines \\ \approx \dfrac{c}{2\sqrt{2}h\sqrt{\varepsilon_r - 1}} = \dfrac{106\ [GHz]}{h[mm]\sqrt{\varepsilon_r - 1}} \quad for\ narrow\ lines\ and\ \varepsilon_r > 10 \end{cases}$$

For higher and higher frequencies of operation,

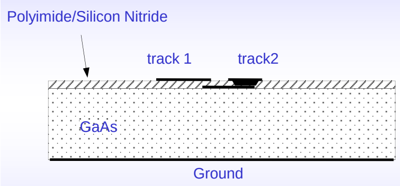
the substrate must be made thinner and thinner !!!!

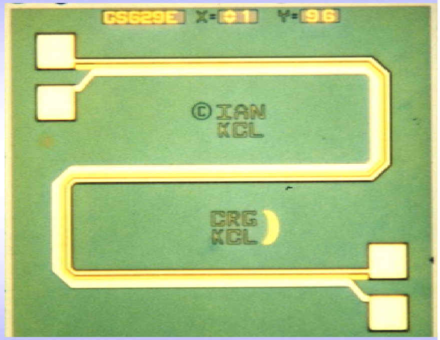
* TM mode limit is usually specified as:-

$$\therefore h_{\max} = \begin{cases} \dfrac{\lambda_o \tan^{-1}(\varepsilon_r)}{\sqrt{2}\pi\sqrt{\varepsilon_r - 1}} \\ \sim \dfrac{\lambda_o}{4\sqrt{\varepsilon_r - 1}} & \text{for wide lines} \\ \sim \dfrac{\lambda_o}{2\sqrt{2}\sqrt{\varepsilon_r - 1}} & \text{for narrow lines and } \varepsilon_r > 10 \end{cases}$$

Radio Frequency Engineering
Lecture #1 Passives

Monolithic Multilayer Microstrip Coupler





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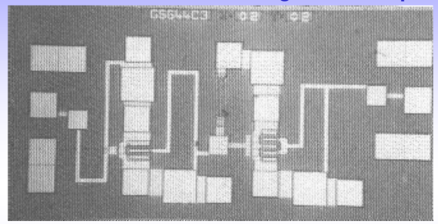
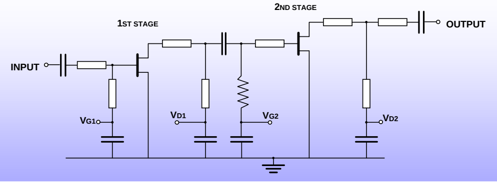
Distributed-element Circuit: 20-40 GHz 2-stage Microstrip MMIC Amplifier





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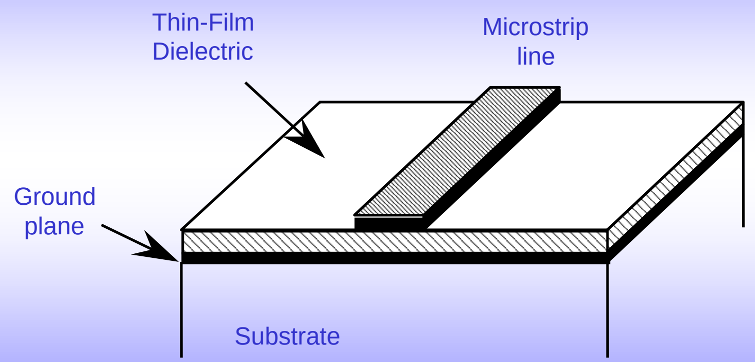
Thin-film Microstrip Transmission (TFMS) Line Medium



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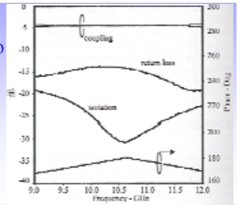
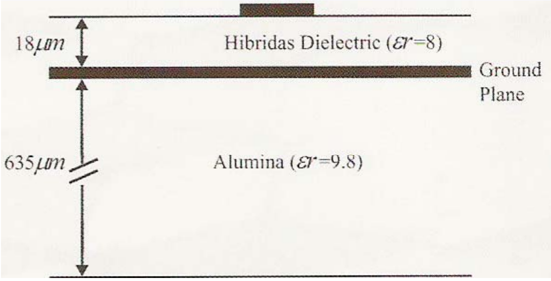
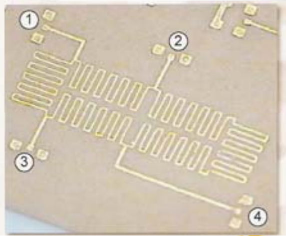
10 GHz TFMS replaces the 620 µm and 270 µm width tracks used to realise conventional 50 Ω and 70 Ω microstrip lines with 30 µm and 15 µm width tracks, respectively.







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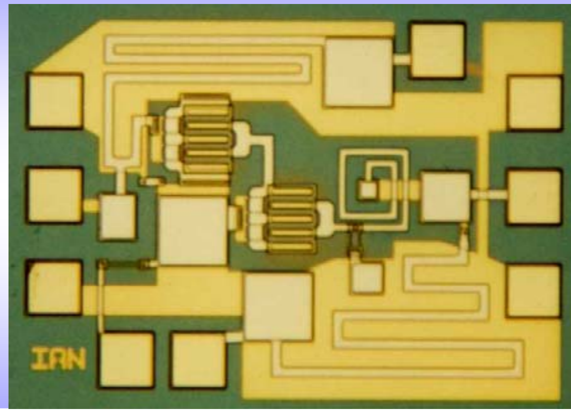
Microphotograph of a Uniplanar (TFMS) MMIC Amplifier



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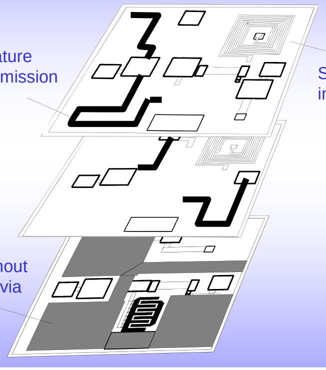
Multilayer MMICs for High Package Density



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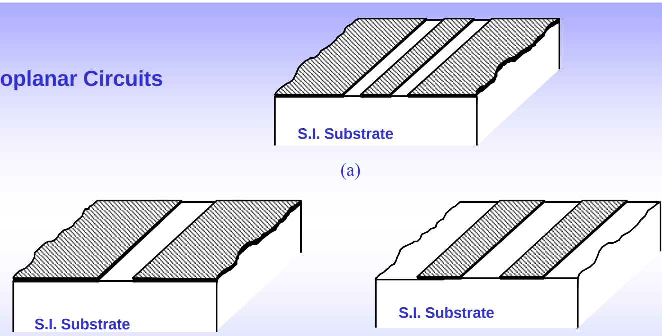
Coplanar Circuits



Ideal transmission lines for uniplanar MICs:
(a) coplanar waveguide (CPW), (b) slotline, (c) coplanar strips (CPSs)

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Principal Advantages of CPW:

1. Devices and components can be grounded without via-holes

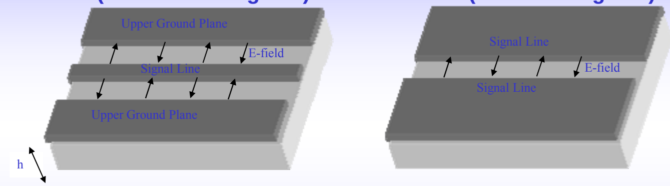
2. It suffers from much less dispersion than microstrip, making it suitable for millimetre-wave circuits

3. A given characteristic impedance can be realized with almost any track width and gap combination

4. A considerable increase in packing density is possible because the ground planes provide shielding between adjacent CPW lines

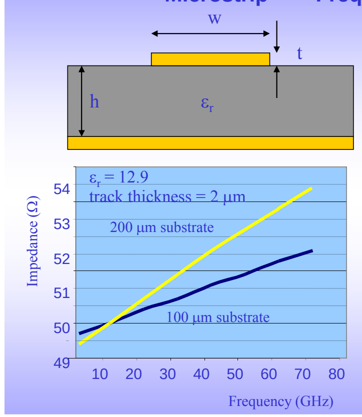
5. With the back-face ground plane removed, lumped-elements exhibit less parasitic capacitance

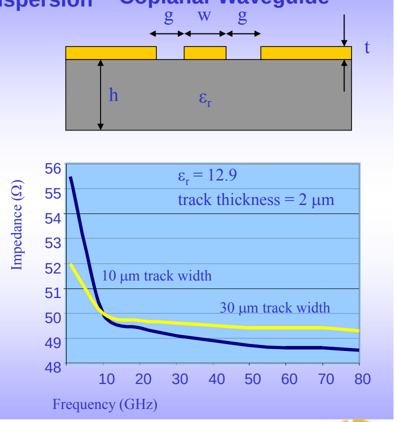
Coplanar Waveguide (unbalanced signals) — Slotline (balanced signals)



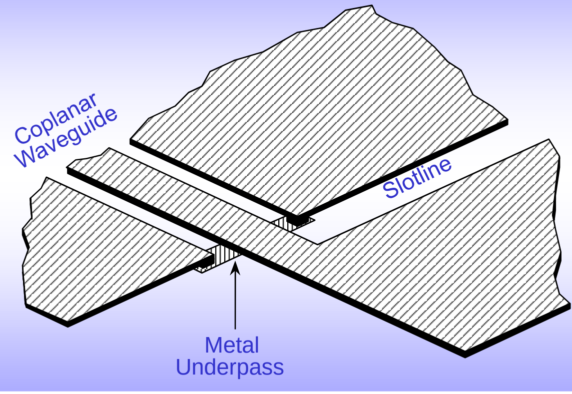
• Slotline transmission lines are used to realise baluns (balanced-to-unbalanced signal transformers), since balanced mixers and push-pull amplifiers require balanced signals from unbalanced sources terminations (to drive two active devices in anti-phase). CPW-to-slotline transitions are commonly used to convert the unbalanced CPW line to the balanced slotline. Here, field-based modelling is essential for designing efficient transitions.

Microstrip — Frequency Dispersion — Coplanar Waveguide

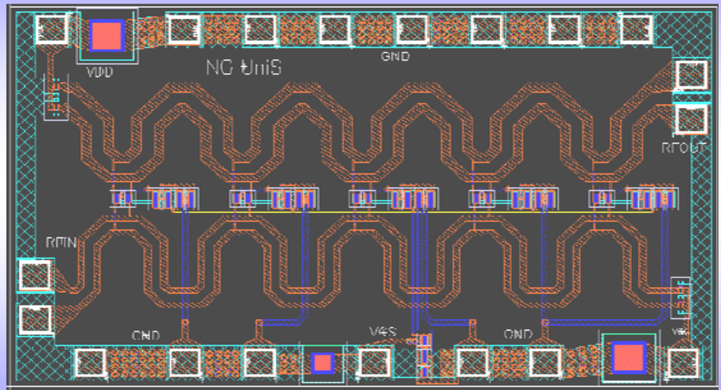




CPW-to-slotline Transition



0.18 μm CMOS CPS Line Distributed Amplifier Design



2.2 mm X 1.3 mm

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Multi-Modeing in Conductor-Backed CPW Lines

With an ideal CPW line, only the pure-CPW (quasi-TEM) mode is considered to propagate. In the case of a grounded-CPW (GCPW) line, where the backside metallization is at the same potential as the two upper-ground planes (through the use of through-substrate vias), a microstrip like mode can also co-exist with the pure-CPW mode. With the conductor-backed CPW (CBCPW) line, where the backside metallization has a floating potential, parallel-plate line (PPL) modes can also co-exist. The significant PPL modes that are associated with CBCPW lines include the fundamental TEM mode (designated TM_0) found at frequencies from DC to infinity and the higher order TMn modes that can only be supported above their cut-off frequency, $f_{cn} \sim nc/(2h\sqrt{\varepsilon_r})$. By inserting a relatively thick dielectric layer (having a lower dielectric constant than that of the substrate), between the substrate and the lower ground plane, the pure CPW mode can be preserved. This is because the capacitance between the upper and lower conductors will be significantly reduced and, therefore, there will be less energy associated with the parasitic modes. Alternatively, the parallel-plate line modes can also be suppressed by reducing the width of the upper-ground planes, resulting in finite ground CPW (or FGC). Finally, in addition to all the modes mentioned so far, the slot-line mode can also propagate if there is insufficient use of air-bridges/underpasses to equalise the potentials at both the upper-ground planes.

PURE-CPW + SLOT-LINE + MICROSTRIP + PARALLEL-PLATE (TEM + TMn)
|-------------CPW-------------|
|------------------GCPW and FGC----------------|
|---CBCPW|---|

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Power Losses

Transmission lines are generally realised using both conductor and dielectric materials, both of which should be chosen to have low loss characteristics. Energy is lost by Joule's heating, multi-modeing and leakage. The former is attributed to ohmic losses, associated with both the conductor and dielectric materials. The second is attributed to the generation of additional unwanted modes that propagate with the desired mode. The latter is attributed to leakage waves that either radiate within the substrate (e.g. dielectric modes and surface wave modes) or out of the substrate (e.g. free space radiation and box modes).

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Losses	Microstrip/CPW/FGC
Dielectric	$\alpha_d \propto$ frequency
Conductor	$\alpha_c \propto Rs \propto$ frequency $^{1/2}$
Radiative	$\propto$ frequency $^{3/2}$

At low frequencies, the conductor losses dominate. If the conductor is removed, as in the case of the dielectric waveguide and optical fibre, the Q-factor increases substantially. However, where there is a conductor, the surface current density should be minimised by spreading the conduction current across as wide an area as possible (e.g. replacing coax in favour of a metal-pipe rectangular waveguide, or CPW in favour of microstrip), so that $|Jc|$ and therefore $|Js|$ are minimised. However, radiation loss in microstrip is more than that in CPW and non-existent in coax and metal-pipe rectangular waveguide.

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For a wide microstrip line, the cut-off frequency at which the dominant leakage mechanism of the TM_0 mode becomes relevant is:

$$f_{C|_{TMo}} = \frac{75}{h\sqrt{\varepsilon_r - 1}} \quad [MHz]$$

For a substrate with either an upper OR lower ground plane, the cut-off frequency at which the dominant leakage mechanism of the TM_1 mode becomes relevant is:

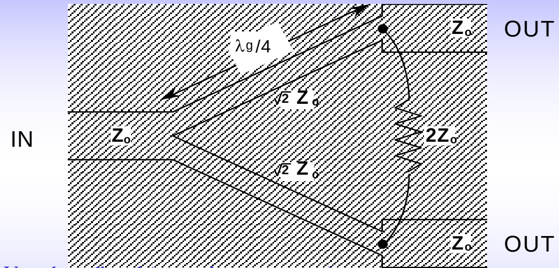
$$f_{C|_{TM1}} = \frac{75}{h\sqrt{\varepsilon_r}} \quad [MHz]$$

For a substrate with either no ground plane OR both upper and lower ground planes, the cut-off frequency at which the dominant leakage mechanism of the TM_1 mode becomes relevant is:

$$f_{C|_{TM1}} = \frac{150}{h\sqrt{\varepsilon_r}} \quad [MHz]$$

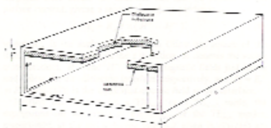
Directional Couplers

Microstrip layout of the Wilkinson power splitter



- Very broadband (more than an octave)
- Requires an isolation (ballast) resistor
- Very simple design
- Meandered lines are possible for lower frequency applications

The TE_{101} box mode, which has the lowest resonant frequency when the height is much less than its width or length, can be effectively damped by placing a dielectric substrate coated with a resistive film onto the lid. Higher order resonant box modes can be suppressed using strategically placed 1 mm thick gum ferrite tiles of electromagnetic absorbers.



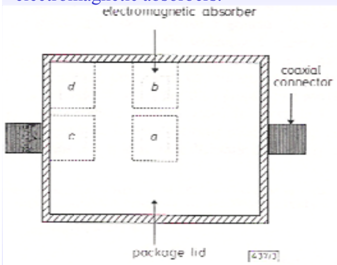


Fig. 3 Electromagnetic absorber placement on back surface lid

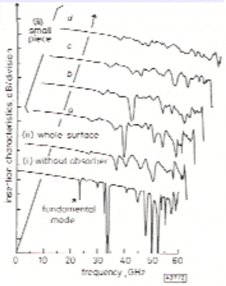


Fig. 2 Experimental results for insertion characteristics a–d: positions shown in Fig. 3

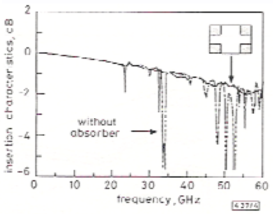
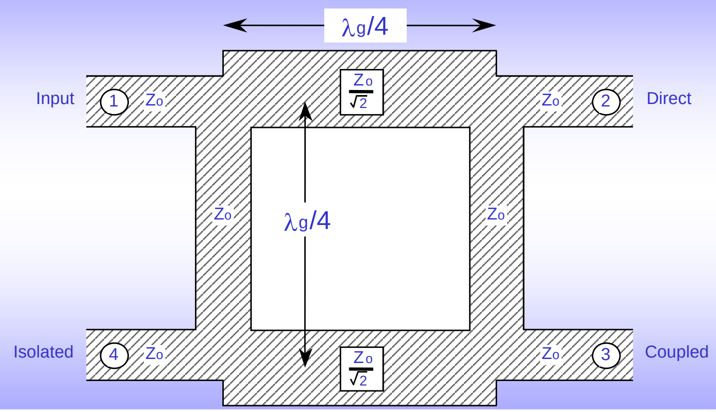


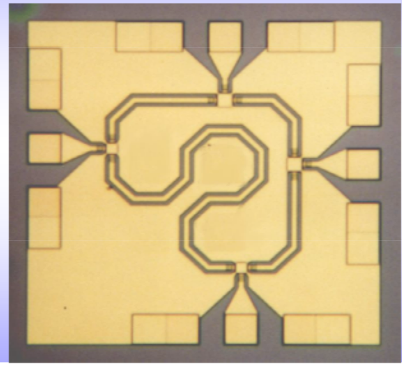
Fig. 4 Insertion characteristics for placement at four co[rners] with those without absorber

The microstrip branch-line coupler

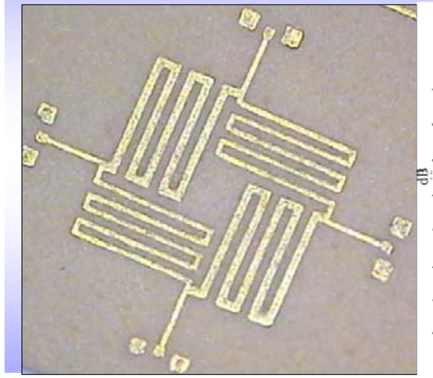


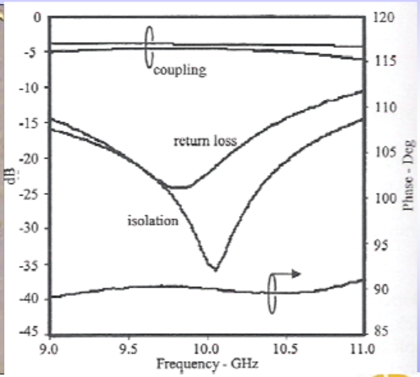
- Works on the interference principle, therefore, narrow fractional bandwidth (15% maximum)
- No bond-wires or isolation resistors required
- Wider tracks make it easier to fabricate and is, therefore, good for lower loss and higher power applications
- Simple design but large
- Meandered lines are possible for lower frequency applications

15 GHz CPW branch-line coupler using the lumped-distributed technique

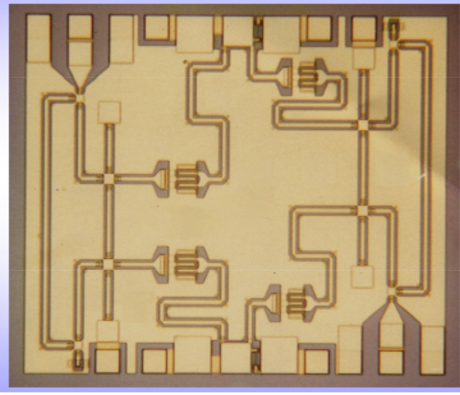


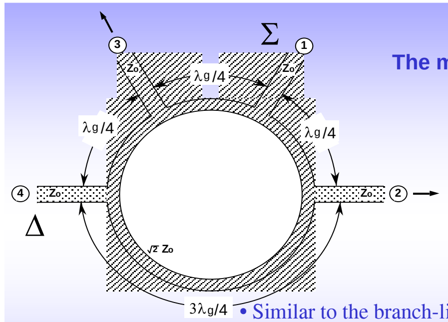
10 GHz TFMS branch-line coupler (approx 1.2 x 1.1 mm^2 with 30 μm and 50 μm tracks)





15 GHz 2-stage balanced MMIC amplifier (employing CPW)





The microstrip rat-race coupler

- Similar to the branch-line coupler with an extra λg/4 length of line
- Works on the interference principle, therefore, narrow fractional bandwidth (15% maximum)
- No bond-wires or isolation resistors required
- Wider tracks make it easier to fabricate and is, therefore, good for

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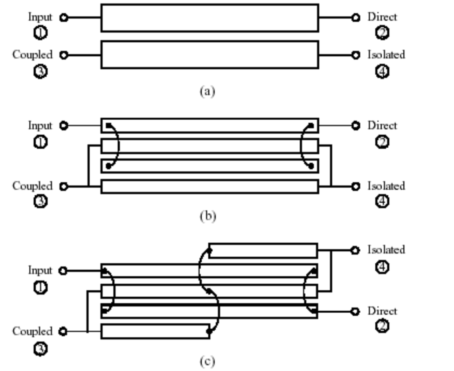

Figure 4.10 Transformation of a basic 2–line directional coupler into a Lange coupler: (a) 2–line directional coupler, (b) unfolded 4–line Lange coupler, and (c) Lange coupler.

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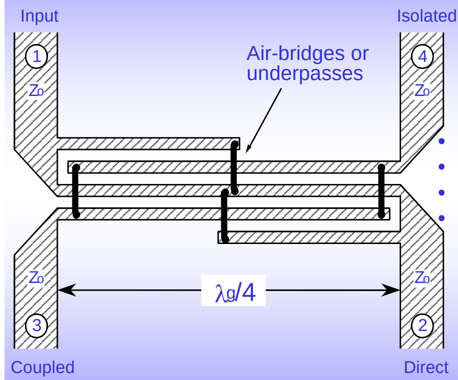
Layout of the Lange coupler



- Very broadband (more than an octave)
- Requires bond-wires
- Narrow tracks can be difficult to pattern
- Bends are possible for lower frequency applications

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Lange Coupler Design

* On a 635 μm thick Alumina substrate the fingers of a Lange coupler are typically 65 μm wide with 53 μm separation

* On a 200 μm thick GaAs substrate the fingers of a Lange coupler are typically 20 μm wide with 10 μm separation

* In determining the exact layout dimensions, frequency dispersion and the effect of any dielectric passivation layers must be taken into account

* The Lange is quite 3D in nature and notoriously hard to model

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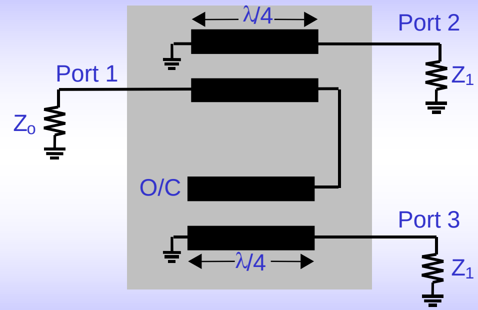
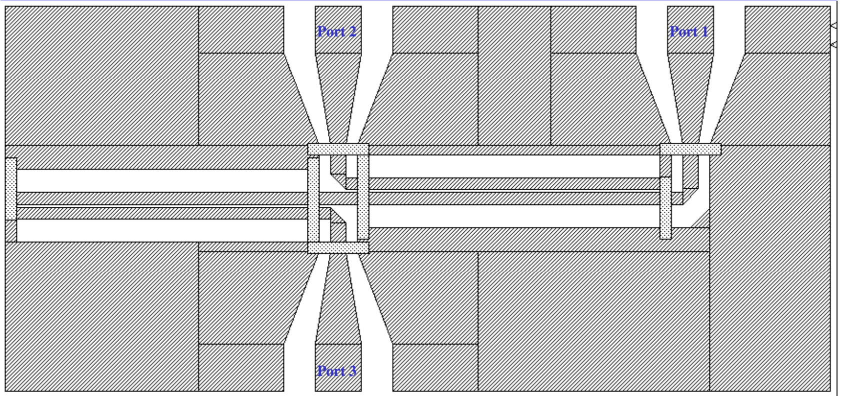
Schematic diagram of planar Marchand balun



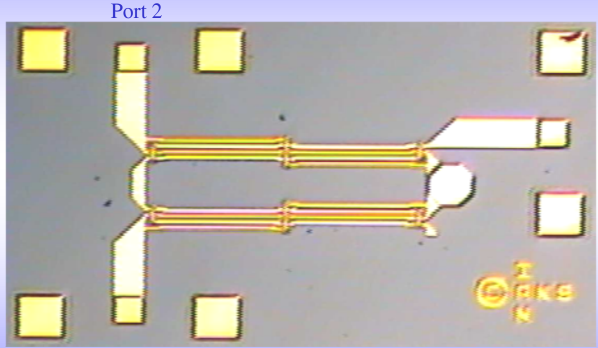
This lossless 3-port network cannot be matched at all ports and two ports are not isolated from one another

Layout of a CPW planar Marchand balun



A Lange coupler planar Marchand balun (0.8 x 1.4 mm^2)

Port 2



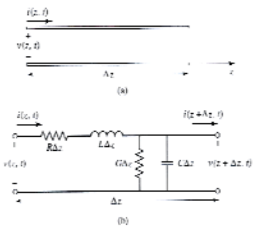
Port 1

Port 3

Basic Transmission Line Theory

Q. Why bother to understand transmission lines?

A. They connect RF sub-systems together, e.g. a transceiver and its antenna. They are also used for impedance matching between circuits and act as resonator elements inside filters and oscillators. Conventional circuit analysis assumes that components are physically much smaller than any wavelength of operation and, therefore, voltages and currents are constant within the individual components. In contrast, transmission line analysis assumes that voltages and currents vary in both magnitude and phase along the length of line. Transmission lines can be analysed using a lumped-element model, but only if the section of line length being considered is very small, i.e. Δz.



L = series self-inductance per unit length [H/m],
represents the H-field associated with both conductors (c.f. $\Psi = LI$)
R = series (conductor loss) resistance per unit length [Ω/m],
represents the ohmic losses in both conductors
C = shunt capacitance per unit length [F/m],
represents the Electric field between conductors (c.f. $Q = CV$)
G = shunt (dielectric loss) conductance per unit length [S/m],
between conductors

$$v(z+\Delta z,t)=v(z,t)-R\Delta z\, i(z,t)-L\Delta z\frac{\partial i(z,t)}{\partial t}$$

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

Therefore, the time-domain form of the Telegraph eqation is:

$$\left.\frac{\Delta v(z,t)}{\Delta z}\right|_{\Delta z\to 0}=\left.\frac{v(z+\Delta z,t)-v(z,t)}{\Delta z}\right|_{\Delta z\to 0}=\frac{\partial v(z,t)}{\partial z}=-\left(R\, i(z,t)+L\frac{\partial i(z,t)}{\partial t}\right)$$

$$\left.\frac{\Delta i(z,t)}{\Delta z}\right|_{\Delta z\to 0}=\left.\frac{i(z+\Delta z,t)-i(z,t)}{\Delta z}\right|_{\Delta z\to 0}=\frac{\partial i(z,t)}{\partial z}=-\left(G\, v(z,t)+C\frac{\partial v(z,t)}{\partial t}\right)$$

For stead-state sinusoidal excitations this simplifies to:

$$\frac{\partial V(z)}{\partial z}=-(R+j\omega L)I(z)$$

$$\frac{\partial I(z)}{\partial z}=-(G+j\omega C)V(z)$$

$$Now,\ Characteristic\ Impedance,\ Zo=\frac{V_+}{I_+}=\frac{V_-}{-I_-}\equiv\frac{R+j\omega L}{\gamma}=\sqrt{\frac{R+j\omega L}{G+j\omega C}}$$

When $R << j\omega L$ and $G << j\omega C$:

$$\alpha=\frac{R}{2}\sqrt{\frac{C}{L}}+\frac{G}{2}\sqrt{\frac{L}{C}}\quad and\quad \beta=\omega\sqrt{LC}$$

When $R = 0$ and $G = 0$ (i.e. purely losses lines):

$$\gamma=0+j\omega\sqrt{LC}\quad and\quad Zo=\sqrt{\frac{L}{C}}=\frac{j\omega L}{\gamma}$$

$$Phase\ Constant,\ \beta=\omega\sqrt{LC}\equiv\frac{2\pi}{\lambda g}=\frac{\omega}{vp}$$

$$Phase\ Velocity,\ vp=\frac{1}{\sqrt{LC}}$$

Differentiating one of these equations and inserting the other equation gives the wave equation for $V(z)$ and $I(z)$:

$$\frac{\partial^2 V(z)}{\partial z^2}-\gamma^2 V(z)=0$$

$$\frac{\partial^2 I(z)}{\partial z^2}-\gamma^2 I(z)=0$$

the travelling wave solutions are now:

$$V(z)=V_+e^{-\gamma z}+V_-e^{+\gamma z}$$

$$I(z)=I_+e^{-\gamma z}+I_-e^{+\gamma z}$$

where, $V_\pm(I_\pm)$ represents voltage (current) waves at z = 0

and, $e^{\mp\gamma z}$ represents wave propagation in the ± z direction

and the propagation constant, $\gamma=\alpha+j\beta\equiv\sqrt{(R+j\omega L)(G+j\omega C)}$

Characteristic Impedance

It has been found that the losses in coaxial cables are at a minimum when Zo ~ 75 Ω. For this reason, 75 Ω coaxial cables are used as standard for video distribution systems, in order to minimise the attenuation of low amplitude video signals. You only need to look at the RF cables feeding your TV, video recorder and associated antenna. Also, in the video industry, the measurement reference impedance is also taken as 75 Ω.

It has also been found that power transmission in a coaxial cable is at a maximum when Zo ~ 30 Ω. For general RF applications, minimal loss and maximum power are of equal importance. For this reason, the RF measurement reference impedance has been standardised to Zo = 50 Ω. This represents average of both the geometric sqrt(75 x 30) = 47.4 Ω and arithmetic (75 + 30)/2 = 52.5 Ω averages.

This explains why laboratory RF signal generators and spectrum analysers have Zo = 50 Ω connectors and use Zo = 50 Ω coaxial cables.

For example, the common RG-58 coaxial cable is filled with solid polyethylene, having: Dk = 2.3, L = 250 nH/m and C = 100 pF/m. Therefore, Zo = 50 Ω and vp = 2 x 10^8 m/s.

Frequency Dispersion

Guided wavelength, λg, is defined as the distance between two successive points of equal phase on the wave at a fixed instance in time.

Phase velocity of a wave is defined as the speed at which a constant phase point travels down the line.

Frequency dispersion is said to occur when $\beta \neq \omega.constant$. Dispersion can occur when $vp = f(\omega)$, i.e. when the dielectric constant $Dk = f(\omega)$. It can be shown that zero dispersion in a lossy line can also occur, but only when $RC = GL$:

$$\gamma^2 = (R + j\omega L)(G + j\omega C) \quad and \quad RC = GL$$

$$\therefore \alpha(\omega) = \alpha(0) = \sqrt{RG} \neq f(\omega) \quad and \quad \beta = \omega\sqrt{LC}$$

$$also, Group\ Velocity, Vg = \frac{\partial\omega}{\partial\beta} = \frac{1}{\sqrt{LC}} \equiv vp \neq f(\omega)$$

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Reflected Waves

Wherever a physical or electrical discontinuity is found on a transmission line, some of the incident electromagnetic energy is reflected back from it, some is absorbed (due to radiation and/or the propagation of surface waves and the generation of evanescent and higher-order modes) and the rest is transmitted through the discontinuity.

$$Voltage\ Wave\ Reflection\ Coefficient,\ \rho(z) = \frac{V_- e^{+\gamma z}}{V_+ e^{-\gamma z}} = \frac{-I_- e^{+\gamma z}}{I_+ e^{-\gamma z}}$$

$$\therefore \rho(z) = \rho(0)e^{+2\gamma z} \equiv \rho(0)e^{+j2\beta z} \quad for\ a\ lossless\ line$$

$$where,\ \rho(0) = \frac{V_-}{V_+} = \frac{-I_-}{I_+}$$

Note that, ρ(z) goes through 360° when z = λg/2 and NOT when z = λg.

$$Termination\ Impedance,\ Z_T = \frac{V(0)}{I(0)} = \frac{V_+ + V_-}{I_+ + I_-} = \frac{V_+ + V_-}{V_+/Zo - V_-/Zo} = Zo\frac{V_+ + V_-}{V_+ - V_-} = Zo\frac{1+\rho(0)}{1-\rho(0)}$$

$$Normalised\ Termination\ Impedance,\ z_T = \frac{Z_T}{Zo} = \frac{1+\rho(0)}{1-\rho(0)}$$

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Bandwidth

Lossless transmission lines can be represented by an infinite number of lumped series-L/shunt-C sections; the self-consistent proof of which is:

$$Zin = j\omega L + \frac{Zo\frac{1}{j\omega C}}{Zo + \frac{1}{j\omega C}} \equiv Zo$$

$$\therefore Zo = \frac{j\omega L}{2}\left(1 \mp \sqrt{1 - \frac{4}{\omega^2 LC}}\right)$$

$$Zo = \begin{cases} \sqrt{\frac{L}{C}} & when \quad \omega << \omega_c \quad i.e.\ purely\ real \\ Complex & when \quad 0 << \omega < \omega_c \\ j\sqrt{\frac{L}{C}} & when \quad \omega = \omega_c \quad i.e.\ purely\ imaginary \\ Imaginary & when \quad \omega \geq \omega_c \end{cases}$$

$\therefore Cut\text{-}off\ frequency,\ fc = \frac{1}{\pi\sqrt{LC}}$ representing the bandwidth, i.e. when $\frac{4}{\omega^2 LC} = 1$

$$Zin(\omega_c) = jZo(\omega << \omega_c) \quad and \quad |\rho(\omega_c)|^2 = 1 \Rightarrow 0\ dB \quad and \quad |\tau(\omega_c)|^2 = 1 - |\rho(\omega_c)|^2 = 0 \Rightarrow -\infty\ dB$$

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This represents a bilinear transformation, which has the property that circles are mapped onto circles (straight lines being considered circles of infinite radius).

$$Therefore,\ \rho(0) = \frac{Z_T - Zo}{Z_T + Zo} \qquad \left(For\ complex\ Zo,\ \rho(0) = \frac{Z_T - Zo^*}{Z_T + Zo}\right)$$

The voltage and current on the line can now be represented as:

$$V(z) = V_+\left(e^{-\gamma z} + \rho(0)e^{+\gamma z}\right)$$
$$I(z) = I_+\left(e^{-\gamma z} - \rho(0)e^{+\gamma z}\right)$$

$$It\ can\ be\ found\ that: V_+ = 0.5(V(0) + ZoI(0)) \quad and \quad V_- = 0.5(V(0) - ZoI(0))$$

$$\therefore incident\ wave\ power,\ P_+ = \frac{|V_+|^2}{Zo} \quad and \quad reflected\ wave\ power,\ P_- = \frac{|V_-|^2}{Zo}$$

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The voltage (and current) on the line is composed of a superposition of the incident and reflected waves, which create a "standing wave", due to the mismatched load termination (even if the generator is matched to the line). Here, the incident and reflected wave magnitudes alternately cancel and reinforce one another. This standing wave disappears when the line is said to be "matched", i.e. $Z_T = Zo$, and we are left with just a single wave travelling in the $+z$ direction.

For a lossless line, the magnitude of the voltage standing wave on the line is:

$$|V(z)| = |V_+ e^{-\gamma z}\left(1+\rho(0)e^{+2\gamma z}\right)| \Rightarrow |V_+|\left|1+\rho(0)e^{+j2\beta z}\right| = |V_+|\left|1+|\rho(0)|e^{+j(\angle\rho(0)-2\theta)}\right|$$

$$where,\ \rho(0) = |\rho(0)|e^{+j\angle\rho(0)} \quad electrical\ length,\ \theta = \beta l \quad and \quad l = -z$$

$$Therefore,\ |V|_{\max} = |V_+|\left(1+|\rho(0)|\right) \quad and \quad |V|_{\min} = |V_+|\left(1-|\rho(0)|\right)$$

$$Voltage\ Standing\ Wave\ Ratio,\ VSWR = \frac{|V|_{\max}}{|V|_{\min}} = \frac{1+|\rho(0)|}{1-|\rho(0)|}$$

Now, the impedance looking into a transmission line that is terminated with a load Z_T is :

$$Zin = \frac{V(l)}{I(l)} = Zo\frac{\left(e^{+\gamma l}+\rho(0)e^{-\gamma l}\right)}{\left(e^{+\gamma l}-\rho(0)e^{-\gamma l}\right)} = \frac{\left((Z_T+Zo)e^{+\gamma l}+(Z_T-Zo)e^{-\gamma l}\right)}{\left((Z_T+Zo)e^{+\gamma l}-(Z_T-Zo)e^{-\gamma l}\right)}$$

$$Zin = Zo\frac{\left(Z_T(e^{+\gamma l}+e^{-\gamma l})+Zo(e^{+\gamma l}-e^{-\gamma l})\right)}{\left(Zo(e^{+\gamma l}+e^{-\gamma l})+Z_T(e^{+\gamma l}-e^{-\gamma l})\right)}$$

$$Therefore,\ zin = \frac{Zin}{Zo} = \frac{z_T+\tanh(\gamma l)}{1+z_T\tanh(\gamma l)} \Rightarrow \frac{z_T+j\tan\theta}{1+jz_T\tan\theta}\ \text{for a lossless line}$$

If Zo is taken to be purely real, the time-average power flow along the line is:

$$\mathrm{P}(z) = \mathrm{Re}\{V(z)I(z)^*\} \Rightarrow (P_+ - P_-) = P_+\left(1-\frac{P_-}{P_+}\right) = \frac{|V_+|^2}{Zo}\left(1-|\rho(0)|^2\right)$$

If you wish to prove this for yourself, remember:

$AA^* = |A|^2$ *and* $(A + B)^* = (A^* + B^*)$ *and* $(A-A^*) = +j2Im\{A\}$

This shows that, for a lossless transmission line, time-average power flow is independent of the line length and is equal to the incident wave power minus the reflected wave power.

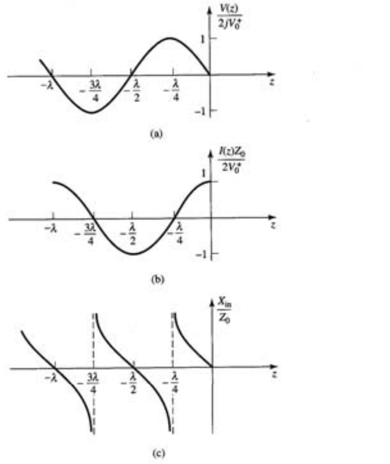
Return Loss, RL = $-10\log(|\rho(0)|^2)$ [dB].

In theory, a pass band circuit should be designed to have a simulated RL < -15 dB, so that the practical measured value would be more like < -10 dB (i.e. < 10% of incident power is reflected back).

- If $l = \lambda g/2$ then $Zin = Z_T$, therefore, no impedance transformation – useful for realising interconnects over a narrow bandwidth.
- If $l = \lambda g/4$ then $Zin = Zo^2/Z_T$, therefore, this is a quarter-wavelength impedance transformer – acts as an impedance inverter over a narrow bandwidth.
- If $Z_T = Zo$ then $Zin = Zo$, therefore, no impedance transformation – useful for realising interconnects over a very wide bandwidth.

If $Z_T = 0$ then $Zin = jZo \tan\theta$ and the impedance is always reactive and periodic along the line, which takes a value from 0 to $+j\infty$ and $-j\infty$ to 0 as l increases from 0 to $\lambda g/4$ and $\lambda g/4$ to $\lambda g/2$.

This is useful for realising any value of "effective" inductance or capacitance over a narrow bandwidth.



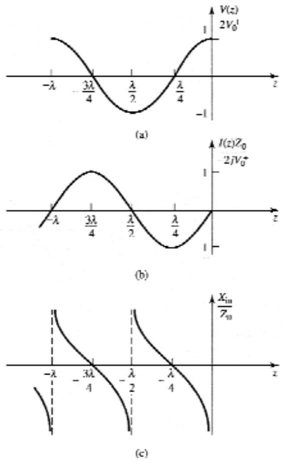
(a) Voltage, (b) current, and (c) impedance (R_{in} = 0 or ∞) variation along a short-circuited transmission line.

Transmission-Line Resonators

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

If $Z_T = \infty$ then $Zin = -jZo \cot\theta$ and the impedance is always reactive and periodic along the line, which takes a value from $-j\infty$ to 0 and 0 to $+j\infty$ as l increases from 0 to $\lambda g/4$ and $\lambda g/4$ to $\lambda g/2$.

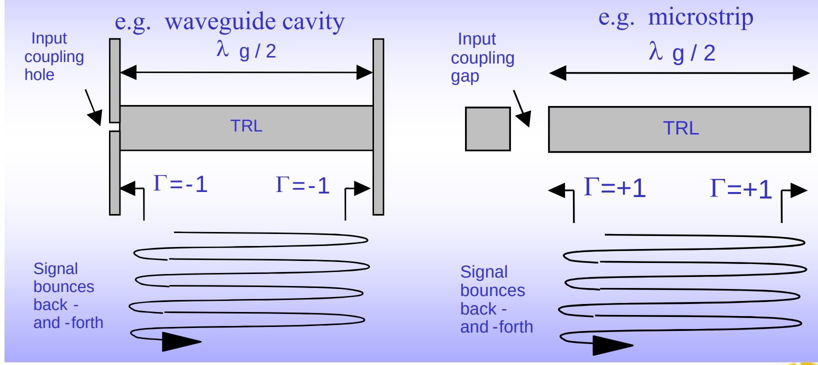
This is useful for realising any value of "effective" capacitance or inductance over a narrow bandwidth.



(a) Voltage, (b) current, and (c) impedance (R_{in} = 0 or ∞) variation along an open-circuited transmission line.

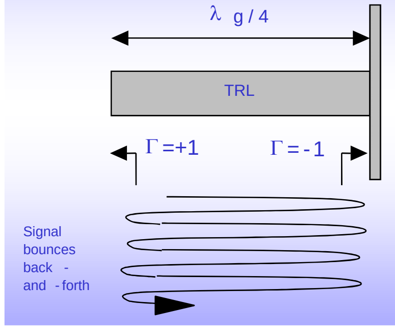
Half-wavelength Resonator, with short circuit terminations

Half-wavelength Resonator, with open circuit terminations

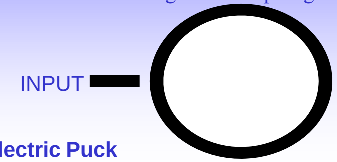


Quarter-wavelength Resonator
- one end open, the other short circuit

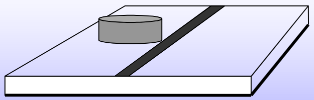
e.g. combline filter



Circular Resonator e.g. microstrip ring



Dielectric Puck Resonator, coupled to microstrip line



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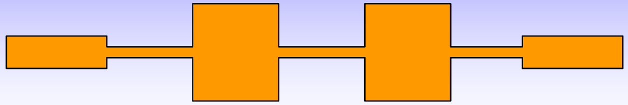
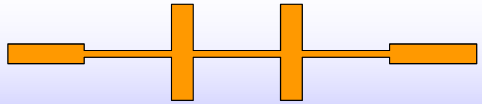
High-Z/Low-Z Low-Pass



- hard to get a big enough line impedance ratio

Low-Pass using shunt stubs



- short open-circuited stubs act as shunt capacitances

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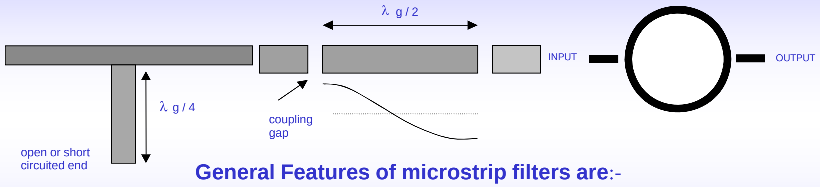
Microstrip Resonators

Quarter-wavelength Resonator | **Half-wavelength Resonator** | **One Wavelength Resonator**



General Features of microstrip filters are:-
- low cost
- easily integrated with active devices
- lossy, low Q, hence performance often not good enough
- low power handling
- Often poor ‘spurious free range’

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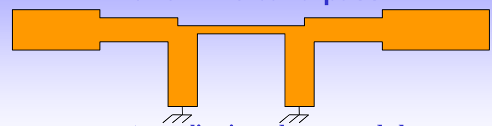
Branch-line band-pass

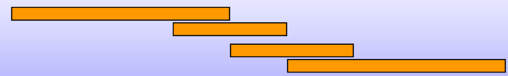


- extreme line impedances needed

End-coupled band-pass

- narrowband only

Parallel-coupled band-pass



- small gaps needed

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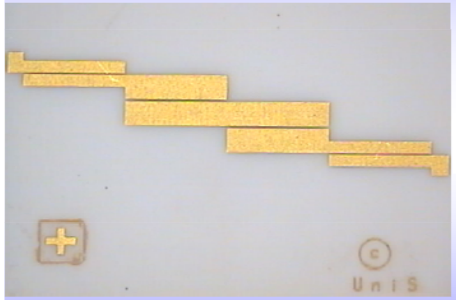
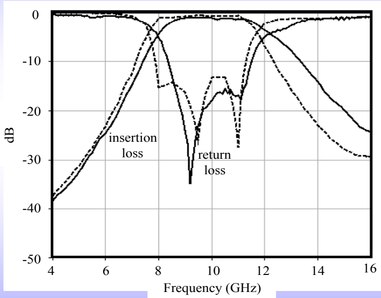
X-Band parallel-coupled Band Pass Filter
Photoimageable Thick Film on Alumina (12 x 12 mm^2)





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Radiating Structures

Maxwell's Equations

1. Faraday Law of Electromagnetic Induction

$$\nabla x\hat{E} = -\frac{\partial \hat{B}}{\partial t} \Rightarrow -j\omega\mu\hat{H} \quad [V/m]$$

where constitutive relationship, $\hat{B}\,[T \text{ or } Wb/m^2] = \mu\,[H/m]\,\hat{H}\,[A/m]$

i.e. electromotive force induced in a closed circuit is proportional to the rate of change in the magnetic flux density threading the circuit.

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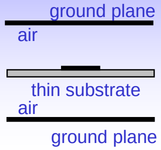
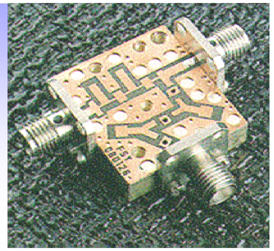
Suspended Substrate Stripline

* Compact Design
* High Reliability
* Easily Fabricated
* 300 MHz to 40 GHz





• Suspended substrate stripline :most of the dielectric surrounding the circuit is removed to increase the structure's unloaded Q.

• lower pass-band loss than conventional stripline filters.

• the lack of dielectric surrounding the circuit makes the filter less sensitive to ambient temperature variations

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2. Ampere's Law of Magnetomotive Force

$$\nabla x\hat{H} = \hat{J}c + \hat{J}_D \Rightarrow \sigma_o\hat{E} + \hat{J}_D \quad [A/m^2]$$

where $\hat{J}c\,[A/m^2] = \sigma_o\,[S/m]\,\hat{E}\,[V/m] \Rightarrow 0$ with in a perfect indulator

i.e. conduction current creates a closed loop of magnetic field. Note that conduction current is simply a surface current density, where the surface is transverse to the direction of current flow (i.e. width x depth). In addition, Maxwell discovered that a magnetic field can also be created by a displacement current. Note that displacement (i.e. electric flux density) is simply a surface charge density.

$$\hat{J}_D = \frac{\partial \hat{D}}{\partial t} = \varepsilon\frac{\partial \hat{E}}{\partial t} \Rightarrow j\omega\hat{E} \quad [A/m^2]$$

where constitutive relationship, $\hat{D}\,[C/m^2] = \varepsilon\,[F/m]\,\hat{E}\,[V/m]$

For example, when a time-varying conduction current flows through the leads of a parallel-plate capacitor then an equal displacement or "wireless" current must also flow between its plates, thus creating a closed loop of magnetic field between the plates.

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3. Gauss's Law: Electric form

$\nabla \cdot \hat{D} = \rho \Rightarrow 0$ with no stored charges $[C/m^3]$

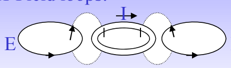
i.e. E-field is created by a stored electric charge. Note that a volume charge density is used.

4. Gauss's Law: Magnetic form

$\nabla \cdot \hat{B} = 0 \quad [Wb/m^3]$

i.e. H-field is not created by a stored magnetic charge. Therefore, it must exist in a closed loop.

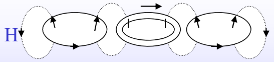
The time-varying H-field, in-turn, creates a time-varying and circulating (i.e. curling) E-field, with lines that encircle the H-Field loops:



This is represented mathematically by Faraday's Law:

$$-\mu \frac{\partial \overline{H}(t)}{\partial t} \rightarrow \nabla x \overline{E}(t)$$

The time-varying E-field, in-turn, creates a time-varying and circulating (i.e. curling) H-field, with lines that encircle the E-Field loops:



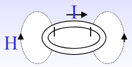
- This is represented mathematically by Maxwell's Law:

$$\varepsilon \frac{\partial \overline{E}(t)}{\partial t} \rightarrow \nabla x \overline{H}(t)$$

- The time-varying H-field, in-turn, creates a time-varying and circulating E-field, etc…..

Propagation of Electromagnetic Waves in Free Space

Propagation can be easily explained by inspection of Maxwell's equations.



Consider a time-varying conduction current (so the free-electron charge is accelerating) flowing around a theoretical loop of lossless wire.

- The conduction current creates a circulating (i.e. curling) H-field, with lines that encircle the current loop. This is represented mathematically by Ampere's Law:

$$\overline{J}_C \rightarrow \nabla x \overline{H}$$

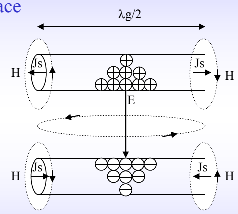
- Since the conduction current is time-varying then so must be the H-field:

$$\overline{J}_C(t) \rightarrow \nabla x \overline{H}(t)$$

Transmitting Dipole Antenna

- A stationary charge will not radiate electromagnetic energy
- A charge moving with constant velocity along a straight wire (i.e. a DC current) creates only a time-invariant H-field and, therefore, no electromagnetic energy will radiate
- An accelerating charge creates a time-variant H-field that, in-turn, creates a time-varying E-field. Therefore, the charge radiates electromagnetic energy. Some example include the following:
  - A charge moving with constant velocity along a curved or bent wire is accelerating
  - When a charge reaches the open end of a wire and reverses direction it accelerates
  - A charge oscillating in simple harmonic motion along a wire periodically accelerates
- Applying a potential across a 2-wire transmission line creates an E-field between the conductors
- E-field lines emanate from a positive charge and terminate at a negative charge. They can also form closed loops.
- The E-field between two lines has a transversal component at the surface of the wires, which displace free electrons in the conductors, giving rise to a H-field
- Since there are no magnetic charges, H-field lines always form closed loops

• It will be shown that an antenna is effectively converting terminated E-field lines (and closed H-field loops) that propagate inside the feed transmission line into closed E- (& H-)field loops that subsequently propagate in free space



• The free electrons associated with the antenna's conductors are analogous to pebbles dropped in water. They are required to excite the waves, but are not needed to sustain the waves

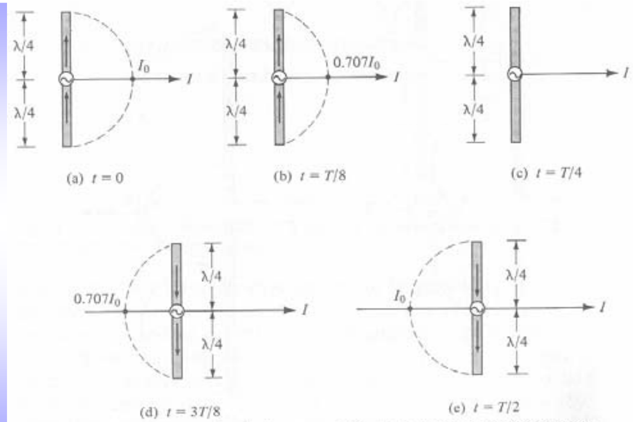
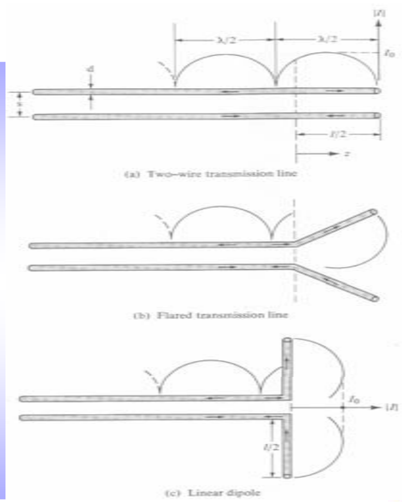
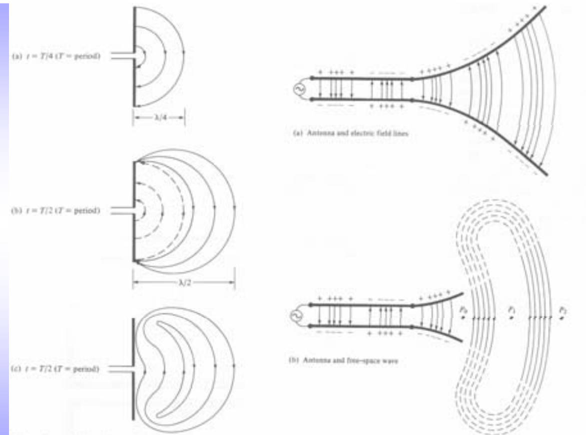


Figure 1.13 Current distribution on a λ/2 wire antenna for different times.

Current distribution on a lossless 2-wire transmission line, flared transmission line and linear dipole





Formation and detachment of electric field lines for short dipole 1.7 Source, transmission line, antenna, and detachment of electric field lines.

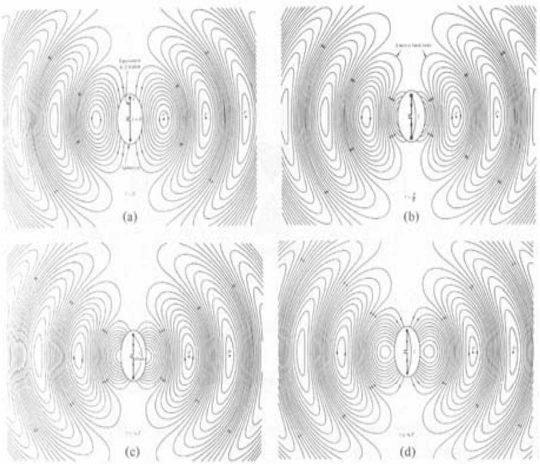


Figure 1.8 Electric field lines of free-space wave for a $\lambda/2$ antenna at $t=0$, $T/8$, $T/4$, and $3T/8$. (SOURCE: J. D. Kraus and K. R. Carver, *Electromagnetics*, 2nd ed.,

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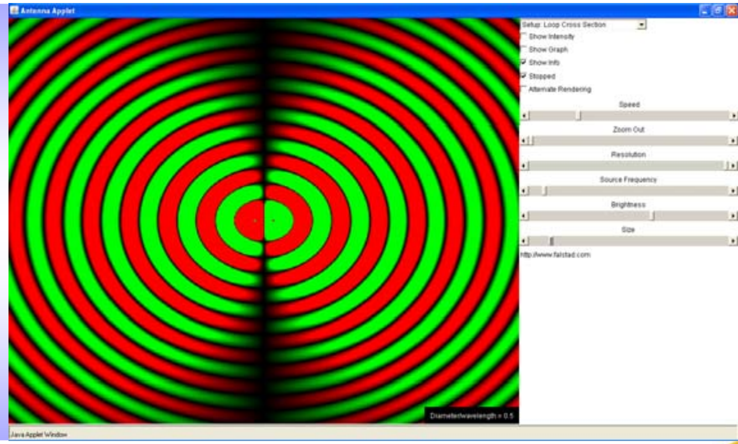

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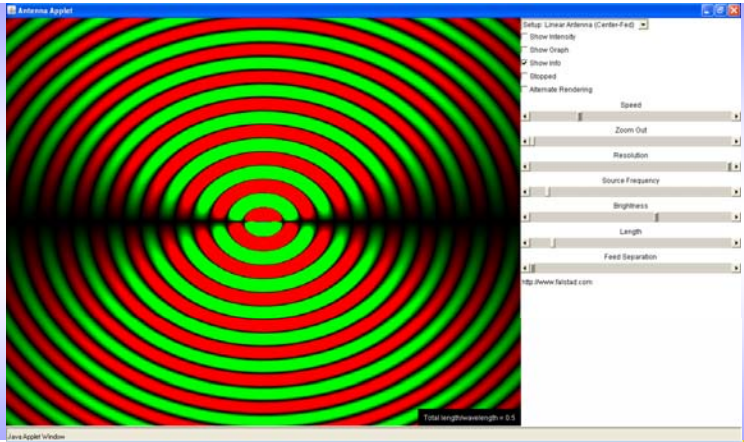

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Plane Waves

- By dropping a pebble in calm water, a circular ripple is created and this grows in size with time (i.e. the wave propagates in 2-dimensions)

- As the radius of the circular ripple increases, its curvature straightens out towards a line. With antenna radiation, the spherical wave approximates a plane wave, whereby the E- & H-field and direction of propagation are all mutually orthogonal (i.e. TEM propagation in free space). At a distance greater than $2(L_{TX}^2 + L_{RX}^2)/\lambda o$ (where L_{*X} = largest spatial dimension of the antenna's aperture), the radiating far field (or Fraunhofer) region exists. What is important is to note that, within this region, the radiation pattern of the antenna is not a function of distance from the antennas. In practice, this minimum distance should be doubled, in order to resolve deep nulls in the radiation pattern (below –25 dB).

- A plane wave has E- and H-fields that are in phase:

$$\overline{E} = \eta \overline{H}$$

where, intrinsic impedance, η, of a lossless medium is a purely real quantity (e.g. with a vacuum)

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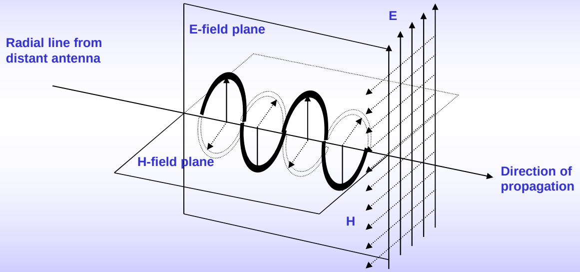
The wavefront is the plane that is mutually orthogonal to the E- & H-planes

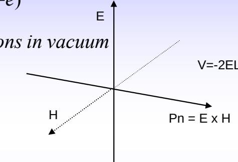


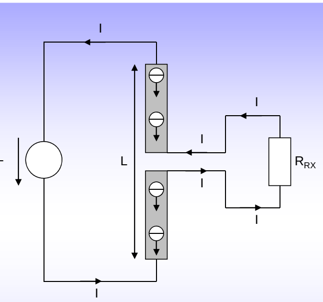
$Work\ (W) = Force\ (F)\ x\ Distance\ (-L)$

$Potential\ Difference\ (V) = \dfrac{W}{(-e)} = -(|\overline{E}| + |\overline{v}x\overline{B}|)L$

$If\ v \equiv c = \dfrac{1}{\sqrt{\varepsilon_o \mu_o}}$ $i.e.\ electrons\ in\ vacuum$

$\therefore c\ B \equiv E\ and\ \therefore V = -2EL$





$If\ v \equiv v_{pc} = \dfrac{\omega}{\beta} = \omega\delta_o = \dfrac{\omega}{\sigma_o Rs_o}\ and\ H = \dfrac{E}{Zs_o} = \dfrac{E}{Rs_o(1+j)}$

$\therefore v_{pc}B \equiv sqrt(2)E\ and\ \therefore V = -2.414EL$

• Antennas obeys the law of reciprocity. For example, the radiation pattern of an antenna under test can be obtained by connecting the signal generator to the reference antenna and then moving this reference antenna around the test antenna (that is connected to a receiver).

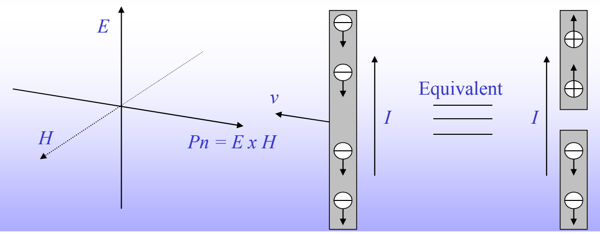
Receiving Dipole Antenna

• A free charge carrier (e.g. electron) will have a force exerted on it by an applied E-field, F_E, and applied H-field, F_H; these forces are described by the Lorentz's force equation:

$$\overline{F} = \overline{F}_E + \overline{F}_H = (-e)(\overline{E} + \overline{v}x\overline{B})$$

(-e) = charge on the electron; v = relative velocity between the electron and orthogonal H-field

• The force contributed by the E- &H-fields are in the same direction with a plane electromagnetic wave and this net force is maximum when the E-field is parallel and H-field is orthogonal to the wire.



Ideal oscillatory electric dipole
Note that:
$I \propto (-e)(+v_d)$
$I \propto (+e)(-v_d)$
v_d = drift velocity

Antenna power gain generally depends on size relative to wavelength

$$\text{Antenna Directivity}, D = 4\pi \frac{\text{Effective Capture Area}, A}{\lambda_o}$$

$$\text{For a parabolic antenna}, A = \pi \left(\frac{d}{2}\right)^2 \text{ and } \therefore D = \left(\frac{\pi d}{\lambda_o}\right)^2$$

$$\text{Antenna Gain} = \text{efficiency} \times \text{Directivity}$$

$$\therefore \text{Gain of an Ideal Parabolic Antenna} = 20\log\left(\frac{\pi d}{\lambda_o}\right) \text{ [dBi]}$$

The largest parabolic antenna in the world is at Arecibo Observatory, Puerto Rico.
With a diameter of 305 m and a 41.6% efficiency, it has a power gain of 70 dBi at 2.38 GHz.

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$$P_{RX} = EIRP \times Spreading\ Loss \times A_{RX}$$

where, Effective Isotropically Radiated Power, EIRP $= P_{TX} \cdot G_{TX}$

and Spreading Loss $= \dfrac{1}{4\pi R^2}$

and Effective Aperture of the Recieving Antenna, $A_{RX} \sim G_{RX} \cdot \dfrac{\lambda_o^2}{4\pi}$

$$\therefore P_{RX} = P_{TX} \cdot \frac{G_{TX} \cdot G_{RX} \cdot \lambda_o^2}{(4\pi R)^2} \quad \textit{Friis' Range Equation}$$

It can be seen from the spreading loss that the received power decreases by:
6 dB when range is doubled and
20 dB for every order of magnitude increase in range.

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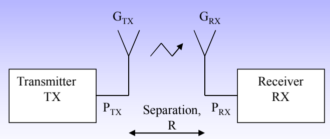
Friis' Link Equation



It is assumed that the separation distance, R, is greater than the combined far-field distance, R_{ff}, for both the transmit and receive antennas (for the gain of both antennas to be known):

$$R_{ff} > \begin{cases} \dfrac{2\left(L_{TX}^2 + L_{RX}^2\right)}{\lambda_o} & \text{for electrically long antennas} \\ 2 \times 2\lambda_o & \text{for electrically short antennas} \end{cases}$$

where, L = Largest Dimention of the Antenna's Aperture

and, λ_o = Free-Space Wavelength

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