Microwave Engineering
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Lecture No. 32
Introduction to Microwave Integrated Circuits (MIC)

We started a new module introduction to microwave integrated circuits, which in short form we call MIC. Now, so far we have seen different types of discrete circuits. We have discussed, for example, impedance matching circuits. We have discussed power dividers. We have discussed directional couplers, filters and also phase shifters.

So, various forms also transistor amplifiers, now when all these circuits or some of these circuits are to be connected together to give, to realize some function or to realize a system, these individual circuits need to be connected and this integration of different circuits leads to integrated circuit and when this is done for the microwave circuits we get microwave integrated circuits.

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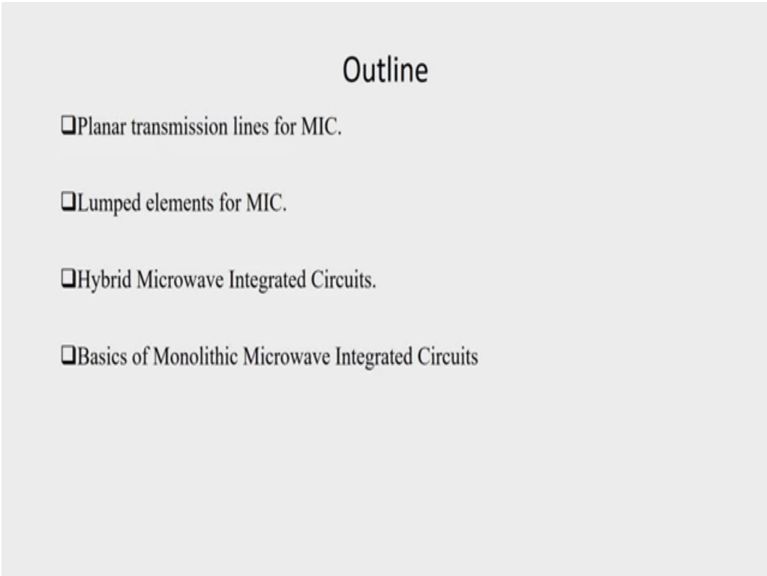

In this module, we discuss planar transmission lines for MIC. Now, we have discussed other forms of transmission lines. We have discussed waveguides, but when it comes to the realization of the integrated circuit, usually planar form of transmission line becomes very useful, and that is why we begin our discussion with planar transmission lines.

And then, we will see the lumped elements for MIC. Here we have the resistors, the capacitors and the inductors in the lumped form that means size very very small compared to the wavelength and which are suitable for integration with other circuitry fabricated using planar technology.

The microwave integrated circuit usually comes in 2 basic forms, hybrid microwave integrated circuit. In hybrid MIC we have the passive components that are fabricated the transmission lines, and some of the passive components are fabricated using planar circuit technology, and then the active components and some of the components which are external are bonded to the remaining circuitry by using metallic connectors. So, this form of microwave circuit is called hybrid microwave circuit, and this is quite a popular form of realizing the system-level functionality using different circuits and integrating them together.

And the most advanced form of microwave circuitry is the monolithic microwave integrated circuit, where the components, transmission lines, everything, they are fabricated on the same substrate.

Now, in hybrid MIC, we can have different substrates for the transmission line, we can have different substrates for other components. Here in monolithic microwave integrated circuit, the same substrate is used where all the components are fabricated, for example gallium arsenide is a popular substrate for realizing this type of MMIC. So, in this module, we are going to discuss some of the basics of this microwave integrated circuit technology, and we start our discussion with the planar form of transmission lines.

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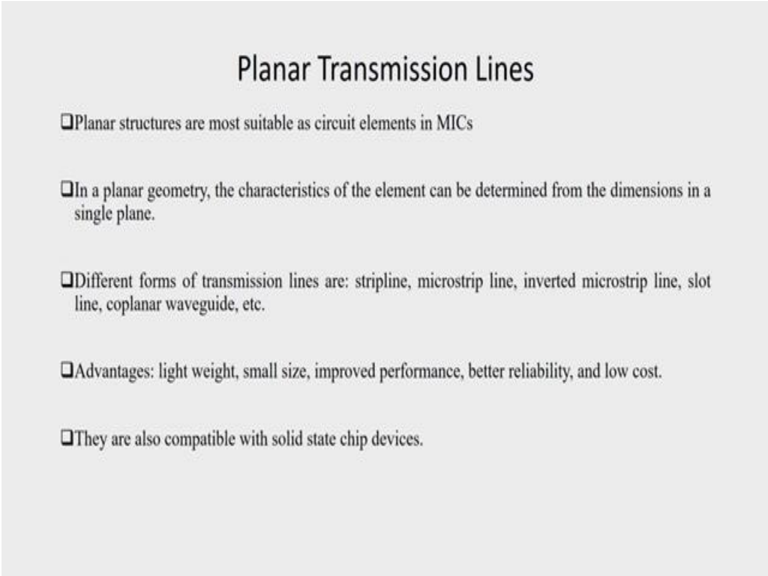

So, there are various types of planar transmission lines, which start with the one which is closer to our co-axial transmission line. This is called a stripline, so planar structures are more suitable as circuit elements in MICs. In planar geometry, the characteristics of the element can be determined from the dimensions in a single plane. And different forms of transmission lines are stripline, microstrip line, inverted microstrip line, slot line, coplanar waveguide, fin line, etcetera.

There are a variety of planar transmission lines and structures which have been developed over the years to satisfy the requirement of circuit design, and this type of transmission lines, they offer different degrees of flexibilities when it comes to realizing a circuit in particular form.

So, planar circuits, planar transmission lines have the advantage of being lightweight, it is smaller in size, provide improved performance and reliability is often better and cost wise also it is low cost. They are also compatible with solid-state chip devices. So, compatibility becomes an issue when we talk of microwave integrated circuits, and in fact the recent trend is that in the low and medium power application, it is the microwave integrated circuits that are becoming popular rather than conventional circuits.

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$$Z_0 = \frac{30\pi}{\sqrt{\epsilon_r}} \frac{b}{W_e + 0.441b}$$

where, W_e is the effective width of the central conductor

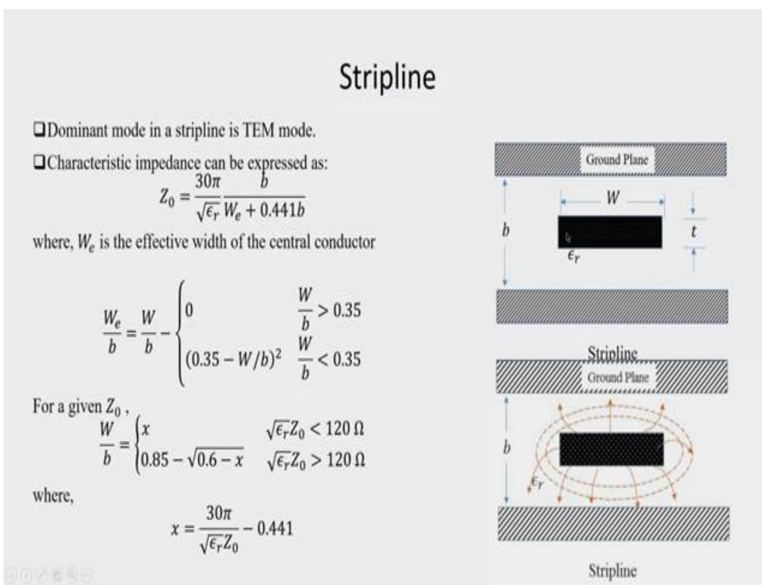
$$\frac{W_e}{b} = \frac{W}{b} - \begin{cases} 0 & \frac{W}{b} > 0.35 \\ (0.35 - W/b)^2 & \frac{W}{b} < 0.35 \end{cases}$$

For a given Z_0 ,

$$\frac{W}{b} = \begin{cases} x & \sqrt{\epsilon_r} Z_0 < 120\ \Omega \\ 0.85 - \sqrt{0.6 - x} & \sqrt{\epsilon_r} Z_0 > 120\ \Omega \end{cases}$$

where,

$$x = \frac{30\pi}{\sqrt{\epsilon_r} Z_0} - 0.441$$



So, we start our discussion with the stripline, so this is the basic structure or geometry of the stripline. And here we can see, a central conductor whose width is denoted by W and thickness is denoted by t. It is between two ground planes, which are separated by a distance b and the region between the ground planes are filled with a dielectric of relative permittivity epsilon r.

Now, the stripline, is a closed structure, the central conductor is being surrounded by ground planes on 2 sides, and therefore the field configuration can be grown as shown in the figure where we have these dotted lines are the magnetic field lines which encircle the central conductor. And the solid lines are the electric field lines which of course have some fringing near the sides of the central conductor. And it will remain confined if the dielectric constant is large, it will remain confined very close to this central conductor.

Now, we can see that the electric and magnetic field lines are perpendicular to each other, and the dominant mode in stripline is TEM mode. The characteristic impedance can be expressed as, so here we are not considering the thickness of the stripline because t is usually very very small. And when t is assumed to be 0, in that case, we can write the characteristic impedance Z naught is equal to 30 pi by root epsilon r b by W e plus 0.441b

Now, this W e is the effective width of the central conductor, and we need to take into account the effective width because of the fringing of fields. W e by b is given as W by b minus if W by b is greater than 0.35 then we did not add any correction factor but if W by b is less than 0.35 in that case the width of the central conductor is much smaller compared to the separation of the two ground planes.

We need to add a correction factor, which is given by 0.35 minus W by b whole square. So, this is when we know W by b epsilon r, in that case we can calculate the characteristic impedance Z naught of the stripline. Otherwise, many a time our requirement is that we know what should be the characteristic impedance, and we need to design the stripline structure for that providing that Z naught.

So, we have given Z naught, we first choose the dielectric constant epsilon r and then we find out this factor x which is 30 pi divided by root epsilon r into Z naught minus 0.441, so once you have calculated this parameter x, then depending upon whether root epsilon r Z naught is less than 120 Ohm or root epsilon r Z naught is greater than 120 Ohm, we can calculate W by b as x or 0.85 minus square root of 0.6 minus x. So, this is the case where we synthesize the stripline for a given Z naught and dielectric constant we have already assumed. So, once we get W by b and then we can design the line if we either specify b or W, the other parameter can be found out.

One problem with the stripline is that the central conductor is between the 2 ground planes. So, when a component is to be connected between the central conductor and the ground, we need to make a hole on the ground plane, and then we get access to the central conductor. So, this is not very convenient for some of the circuits, so you go for another design of the planar transmission line, which is called a microstrip line, where both the conductors, ground planes, and the conductor both are accessible, and this form of line is very popular.

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$$Z_0 = \frac{30}{\sqrt{\epsilon_r}} \ln\left\{1 + \frac{4b - t}{\pi W^t}\left[\frac{8b - t}{\pi W^t} + \sqrt{\left(\frac{8b - t}{\pi W^t}\right)^2 + 6.27}\right]\right\}$$

$$\frac{W^t}{b - t} = \frac{W}{b - t} + \frac{\Delta W}{b - t}$$

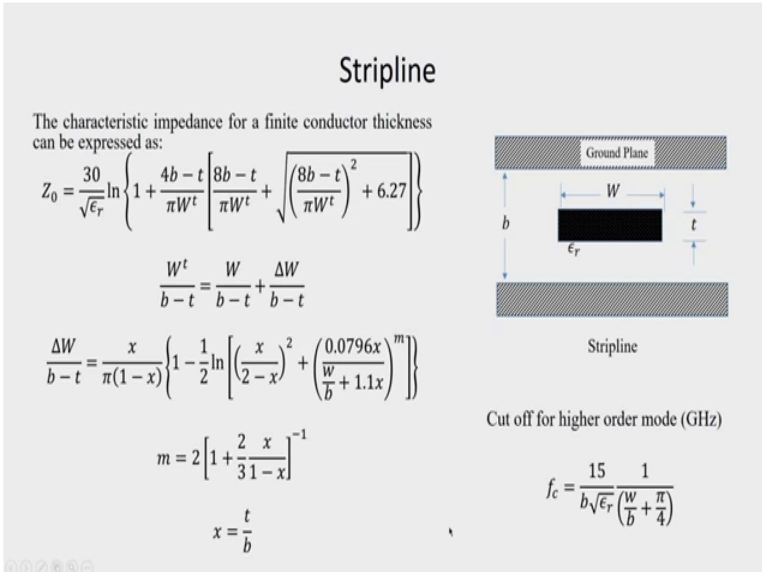
$$\frac{\Delta W}{b - t} = \frac{x}{\pi(1 - x)}\left\{1 - \frac{1}{2}\ln\left[\left(\frac{x}{2 - x}\right)^2 + \left(\frac{0.0796x}{\frac{W}{b} + 1.1x}\right)^m\right]\right\}$$

$$m = 2\left[1 + \frac{2}{3}\frac{x}{1 - x}\right]^{-1}$$

$$x = \frac{t}{b}$$

Cut off for higher order mode (GHz)

$$f_c = \frac{15}{b\sqrt{\epsilon_r}}\frac{1}{\left(\frac{W}{b} + \frac{\pi}{4}\right)}$$



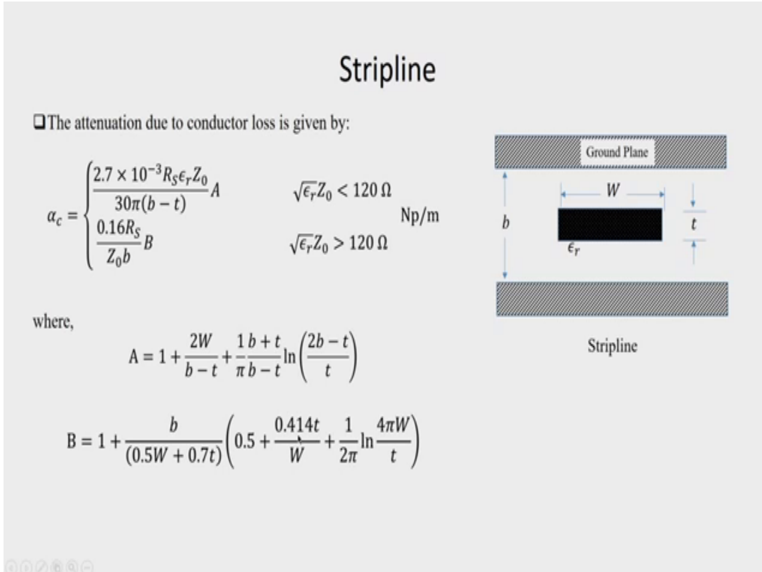
The attenuation due to conductor loss is given by:

$$\alpha_c = \begin{cases} \frac{2.7 \times 10^{-3} R_S \epsilon_r Z_0}{30\pi(b-t)} A & \sqrt{\epsilon_r} Z_0 < 120\ \Omega \\ \frac{0.16 R_S}{Z_0 b} B & \sqrt{\epsilon_r} Z_0 > 120\ \Omega \end{cases} \quad \text{Np/m}$$

where,

$$\mathrm{A} = 1 + \frac{2W}{b-t} + \frac{1}{\pi}\frac{b+t}{b-t}\ln\left(\frac{2b-t}{t}\right)$$

$$\mathrm{B} = 1 + \frac{b}{(0.5W + 0.7t)}\left(0.5 + \frac{0.414t}{W} + \frac{1}{2\pi}\ln\frac{4\pi W}{t}\right)$$



So, we have seen the characteristic impedance for the stripline when the conductor thickness is 0, let us now consider the characteristic impedance of stripline when we have finite conductor thickness. So, when we have finite conductor thickness Z naught needs to be calculated using the relationship as shown. And here we start with this parameter t by b and then we can calculate m, delta W divided by b minus t and then finally these equations may be substituted to calculate Z naught. Please note that this type of relationship, these are of an empirical relation and obtained through a series of experimentation with different dielectric constant, different thickness of the central conductor, different separation b, and then these closed-form relations are obtained through car fitting.

Now, stripline, as we have mentioned it, carries TEM wave. Now, if the frequency of the operation is increased above certain cut off frequency, this stripline can support higher-order modes. So, the cut off for this higher-order mode can be calculated for a given line using this expression f_c is equal to 15 divided by b root epsilon r 1 by W by b plus pi by 4. So, if we want to avoid excitation of higher-order modes, we must design the stripline such that it operates below the cut off frequency.

Now, in a stripline, the attenuation of the signal occurs because of the losses in the conductors as well as the losses in the dielectric. And this conductor and dielectric losses have been estimated, and from there we can get the expression for attenuation constant. So, when the losses due to conductor loss we get alpha c, the attenuation constant because of conductor loss given by this expression and for a given transmission line, stripline configurations with values of W t epsilon r and b known, attenuation constant alpha can be calculated. We need additional information about Rs the sheet resistance.

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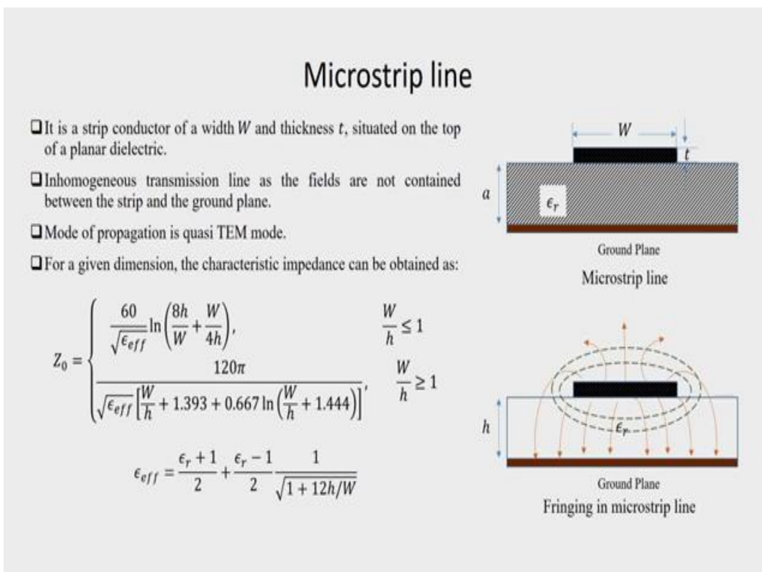
$$Z_0 = \begin{cases} \dfrac{60}{\sqrt{\epsilon_{eff}}} \ln\left(\dfrac{8h}{W} + \dfrac{W}{4h}\right), & \dfrac{W}{h} \leq 1 \\ \dfrac{120\pi}{\sqrt{\epsilon_{eff}}\left[\dfrac{W}{h} + 1.393 + 0.667 \ln\left(\dfrac{W}{h} + 1.444\right)\right]}, & \dfrac{W}{h} \geq 1 \end{cases}$$

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12h/W}}$$



Similarly, the attenuation constant we have shown is only considering the losses in the conductor. The dielectric losses have been neglected. But if the dielectric is (())(22:42), we can find out the power loss because of the loss in dielectric. Stripline has a major disadvantage is that the central conductor it is in between 2 ground planes and if a component needs to be connected between central conductor and ground plane, the central conductor has to be made accessible by drilling holes on the ground plane which may disturb the functioning of the stripline. We consider a more popular form of planar transmission line where both the conductors are accessible, and this form of transmission line is called microstrip transmission line.

As we show here, a microstrip transmission line has a ground plane. It has a conductor, a strip of width W, and thickness t separated from the ground plane by a dielectric material, having dielectric constant epsilon r and thickness h. The propagation in microstrip is called quasi TEM because unlike stripline here dielectric is present only on one side, the other side is here, and therefore the fringing of the field takes place. The transmission is close to TEM but not purely TEM, and this type of propagating mode are called quasi TEM.

So, it consists of a conductor of width W, thickness t situated on the top of a planar dielectric. It is an inhomogeneous transmission line. Fields are not contained between strip and ground plane. Fringing of fields occur. The mode of propagation is called quasi TEM. And for a given dimension, the characteristic impedance can be obtained as, as we can see the characteristic impedance can be obtained as Z naught is equal to 60 by under root eta effective log of ln 8h by W plus W by 4h.

Now, this equation is used when W by h is very very less compared to 1, that means we have a very thin strip, and the microstrip line is very, very narrow. Whereas, when W by h is greater than 1, then we have Z naught is equal to 120 pi under root eta effective W by h plus 1.393 plus 0.667 ln W by h plus 1.444, where this eta effective it depends upon epsilon r and h by W. Why we are talking of eta effective? Because here the strip is not fully inside a single dielectric media.

So, the first approximation of the dielectric constant is epsilon r plus 1 divided by 2. And this is the correction factor which is applied, and please note that these equations are derived through car fitting of experimentally gathered data, and this form of characteristic impedance is, these expressions are quite standardized now.

Now, this equation what we have seen. It is the analysis equation. This means given W, h, and epsilon r, please note that we are assuming the thickness to be very small, to be neglected, t is almost 0, and given W, h and epsilon r we can find out, what will be the characteristic impedance of the line. But in certain cases we may have to find out the dimensions of these lines when we have specified the characteristic impedance.

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For a given Z_0 and ϵ_r, the $\frac{W}{h}$ can be obtained as

$$\frac{W}{h} = \begin{cases} \dfrac{8e^4}{e^{2A} - 2}, & \dfrac{W}{h} < 2 \\ \dfrac{2}{\pi}\left[B - 1 - \ln(2B - 1) + \dfrac{\epsilon_r - 1}{2\epsilon_r}\left\{\ln(B - 1) + 0.39 - \dfrac{0.61}{\epsilon_r}\right\}\right], & \dfrac{W}{h} > 2 \end{cases}$$

where,

$$A = \frac{Z_0}{60}\sqrt{\frac{\epsilon_r + 1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1}\left(0.23 + \frac{0.11}{\epsilon_r}\right)$$

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}}$$

$$Z_0 = \frac{1}{c\sqrt{CC_a}}$$

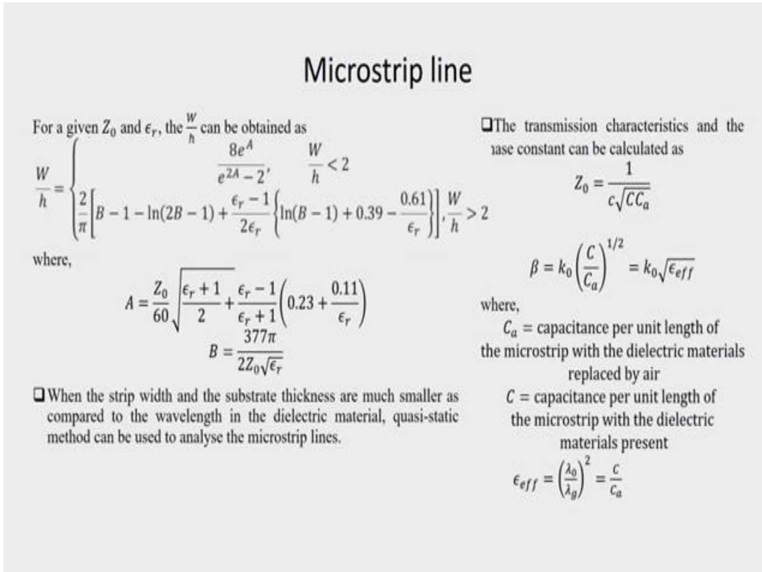
$$\beta = k_0\left(\frac{C}{C_a}\right)^{1/2} = k_0\sqrt{\epsilon_{eff}}$$

where,

C_a = capacitance per unit length of the microstrip with the dielectric materials replaced by air

C = capacitance per unit length of the microstrip with the dielectric materials present

$$\epsilon_{eff} = \left(\frac{\lambda_0}{\lambda_g}\right)^2 = \frac{C}{C_a}$$



So, for a given Z naught, suppose I want to design a microstrip line which gives me 50 Ohm characteristic impedance, in this case and suppose I have been specified the epsilon r, in this case, we need to calculate W by h and for this synthesis of transmission line W by h can be calculated once again based on the fact that whether the line is very narrow or the line is wide line. W by h greater than 2 means width of the strip is 2 times the dielectric thickness whereas W by h less than 2, this is a case where the strip is very narrow, and we can carry out the design of such lines from the given information about Z naught and epsilon r and calculate W by h.

So, we can see that we need to calculate these two parameters first A and B and once we have this values of A and B calculated, then either we can find out W by h, for W by h less than 2 or we can calculate W by h when W by h is greater than 2. When the strip width and the substrate thickness are much smaller compared to the wavelength in the dielectric material, the quasi-static method can be used to analyze the microstrip line. The transmission characteristics and the phase constant can be calculated as Z naught is equal to 1 by c, the velocity of light, root CCa and we have beta is equal to k naught square root of C by Ca, which is equal to k naught root epsilon effective.

So, Ca is the capacitance per unit length of the microstrip with the dielectric material replaced by air and C is the capacitance per unit length of the microstrip with the dielectric material present. And eta effective is given by lambda by lambda g whole square, which is equal to C by Ca.

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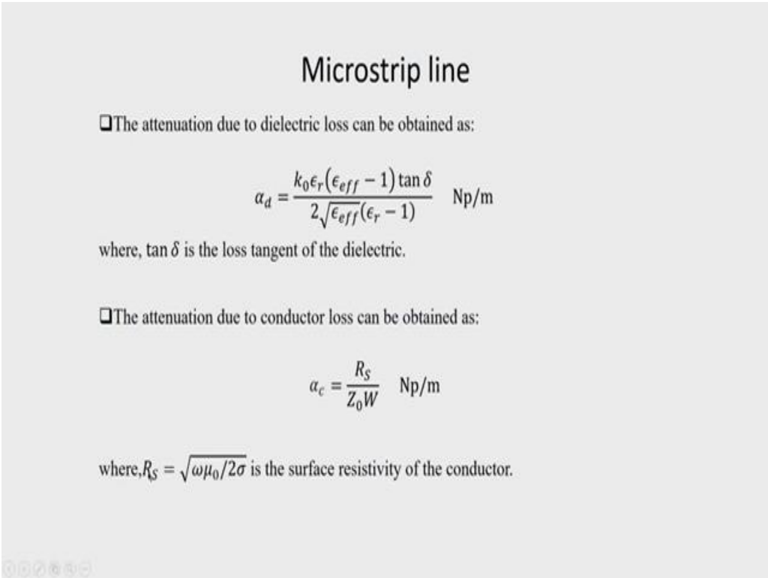
$$\alpha_d = \frac{k_0\epsilon_r(\epsilon_{eff}-1)\tan\delta}{2\sqrt{\epsilon_{eff}}(\epsilon_r-1)} \quad \text{Np/m}$$

$$\alpha_c = \frac{R_S}{Z_0 W} \quad \text{Np/m}$$



So, the attenuation due to dielectric loss can be obtained from the effective epsilon, and this is denoted by alpha d. And we can also estimate the attenuation due to conductor loss, and this can be written as alpha c is equal to Rs divided by Z naught W where Rs is the sheet resistance or the surface resistivity of the conductor.

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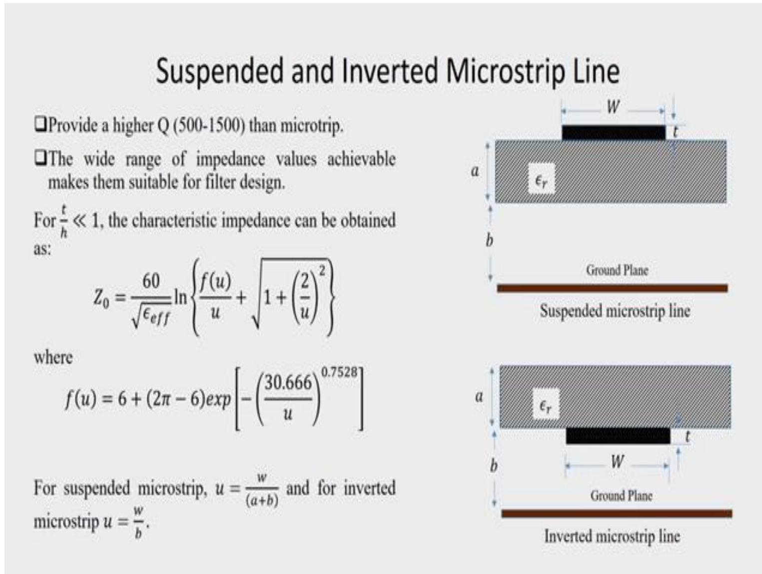
For $\frac{t}{h} \ll 1$, the characteristic impedance can be obtained as:

$$Z_0 = \frac{60}{\sqrt{\epsilon_{eff}}}\ln\left\{\frac{f(u)}{u} + \sqrt{1+\left(\frac{2}{u}\right)^2}\right\}$$

where

$$f(u) = 6 + (2\pi - 6)exp\left[-\left(\frac{30.666}{u}\right)^{0.7528}\right]$$

For suspended microstrip, $u = \frac{w}{(a+b)}$ and for inverted microstrip $u = \frac{w}{b}$.



For suspended microstrip,

$$\sqrt{\epsilon_{eff}} = \left[1 + \frac{a}{b}\left(a_1 - b_1 \ln\frac{W}{b}\right)\left(\frac{1}{\sqrt{\epsilon_r}} - 1\right)\right]^{-1}$$

where,

$$a_1 = \left(0.8621 - 0.1251 \ln\frac{a}{b}\right)^4$$

$$b_1 = \left(0.4986 - 0.1397 \ln\frac{a}{b}\right)^4$$

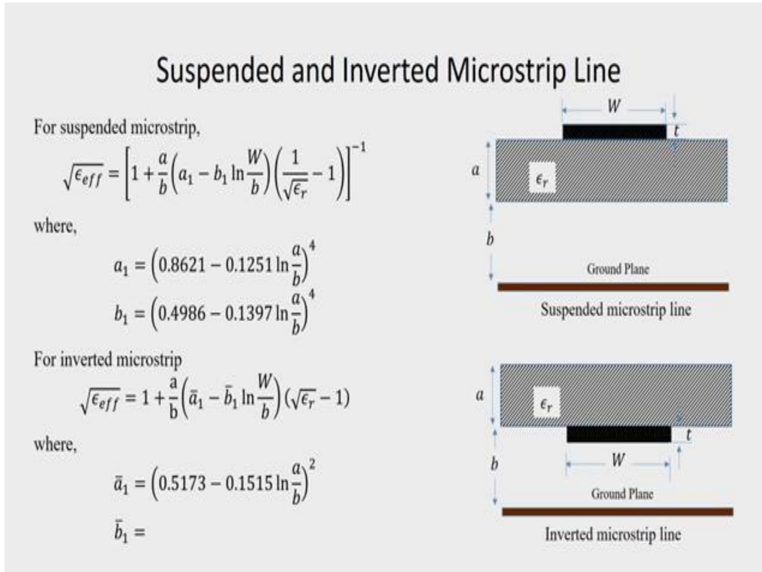
For inverted microstrip

$$\sqrt{\epsilon_{eff}} = 1 + \frac{\mathrm{a}}{\mathrm{b}}\left(\bar{a}_1 - \bar{b}_1 \ln\frac{W}{b}\right)\left(\sqrt{\epsilon_r} - 1\right)$$

where,

$$\bar{a}_1 = \left(0.5173 - 0.1515 \ln\frac{a}{b}\right)^2$$

$$\bar{b}_1 = \left(0.4986 - 0.1397 \ln\frac{a}{b}\right)^2$$



So, we have discussed the basic microstrip form, which is very popular, and this form of microstrip is useful in design of different circuits, antennas and also for connecting the various components together. The other forms of planar transmission lines are suspended and inverted microstrip line, where we have in case of suspended we have ground plane and the strip along with the dielectric there is a gap between ground plane and the dielectric, and this is an air-filled region, and we can see that the strip is suspended. Inverted microstrip line has a ground plane on top of the strip, while the strip when we consider from the strip to the ground plane, we have air in between.

So, this type of configuration is used because they provide higher Q than microstrip. So, when we have requirement of higher Q we can go for this type of configuration. The wide range of impedance values achievable makes them suitable for filter design. So, suspended microstrip line and inverted microstrip lines, they are used in filter design.

The characteristic impedance for t by h much less than 1, the characteristic impedance can be found as Z naught is equal to 60 by square root effective log ln of f of u by u plus 1 plus root 2 by u whole square. Where we define f of u and u as follows, u is W divided by a plus b, when it is a suspended microstrip, and this is equal to W by b when it is in inverted microstrip.

And then we can find out substituting u. We can find out this function f u, and then when f u by u is substituted in Z naught, we get the desired expression for Z naught. For suspended microstrip we calculate eta effective as before, and then we calculate the parameter a1 and b1. And then for the inverted microstrip we have epsilon effective given by this.

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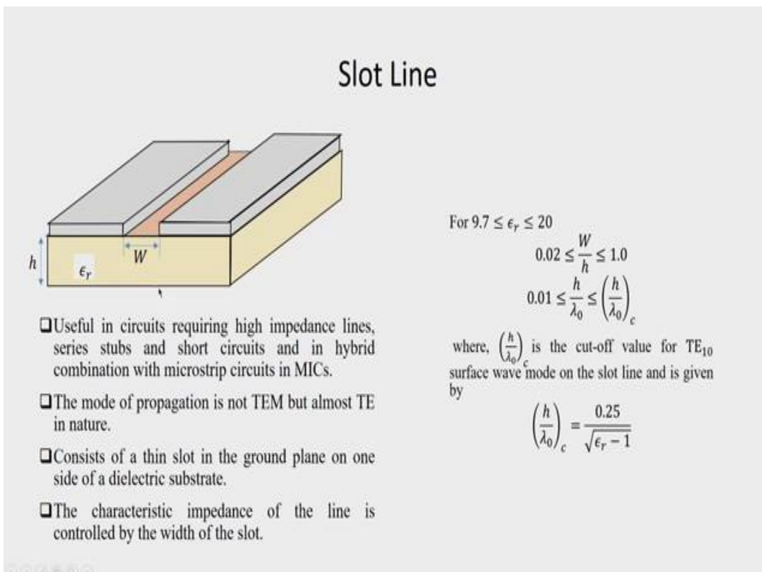
For $9.7 \leq \epsilon_r \leq 20$

$$0.02 \leq \frac{W}{h} \leq 1.0$$

$$0.01 \leq \frac{h}{\lambda_0} \leq \left(\frac{h}{\lambda_0}\right)_c$$

where, $\left(\frac{h}{\lambda_0}\right)_c$ is the cut-off value for TE_{10} surface wave mode on the slot line and is given by

$$\left(\frac{h}{\lambda_0}\right)_c = \frac{0.25}{\sqrt{\epsilon_r - 1}}$$



Another form of the transmission line is the slot line, and it is a useful circuit when high impedance is required. And the mode of propagation is closer to transverse electric or TE rather than TEM mode. And consists of a thin slot on the ground plane of one side of a dielectric substrate. So, we have dielectric substrate, we have ground plane, and a thin slot is being cut.

Now, this type of line usually use the higher value of dielectric constant, and for dielectric constant greater than 9.7 and less than 20, we can define W by h can be from 0.02 to 1. And h by lambda naught, lambda naught is the free space wavelength, it is given by 0.25 divided by epsilon r minus 1. So, we can find out W by h from this information.

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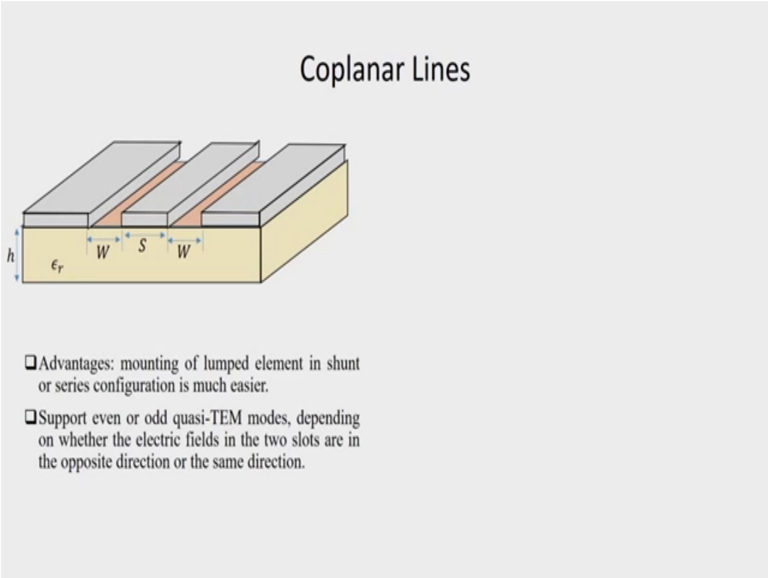

Another form of the planar transmission line, it is the coplanar line, and here the advantage is that the ground plane and the signal conductor, both are on the same plane. And mounting of lumped element in shunt or series configuration to realize integrated circuits become very easy. Suppose this support even and odd quasi modes, depending on whether electric field in the two slots is in the opposite direction or in the same direction.

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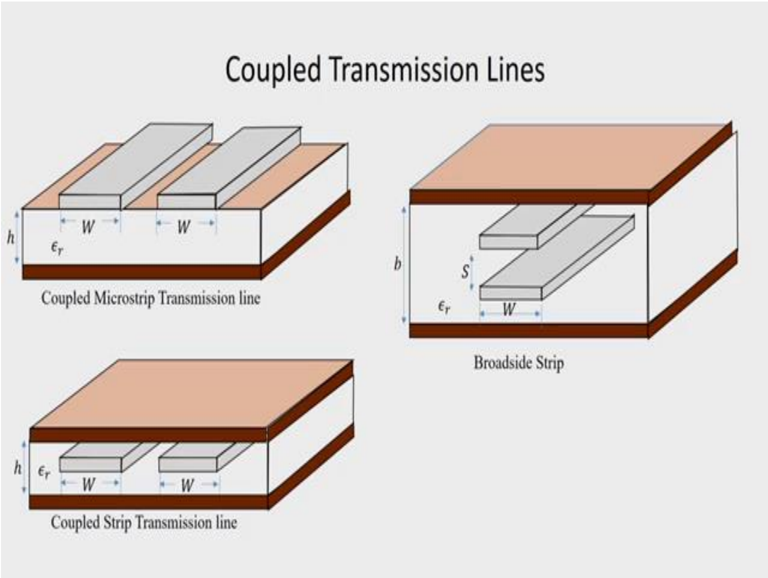

We go to other forms of a transmission line, which is a coupled transmission line and here this type of line is extensively used for designing filters. Here we have 2 microstrip lines, they are separated by a distance, and they get coupled. We also have another form of line, which is a coupled stripline where these 2 conductors, central conductors are side by side, and we have ground plane on top and bottom.

Usually these type of coupled strip lines or coupled microstrip lines, they have now side by side overlapping of the central transmission conductor. Whereas we can have a broadside overlap and the strips may be more than 1 strip may be there, and we have 2 strips, these are stacked one above the other.

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$$Z_0 = \frac{30\pi}{\sqrt{\epsilon_{eff}}}\frac{K(k')}{K(k)}$$

K represents a complete elliptic function of the first kind with K' as its complementary function.

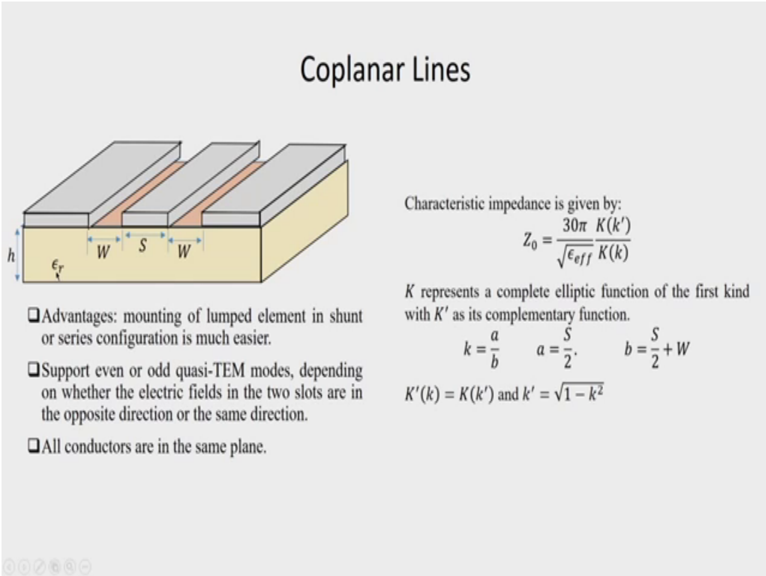
$$k = \frac{a}{b} \qquad a = \frac{S}{2}. \qquad b = \frac{S}{2} + W$$

$K'(k) = K(k')$ and $k' = \sqrt{1-k^2}$

$$\epsilon_{eff} = 1 + \frac{\epsilon_r - 1}{2}\frac{K(k')K(k_1)}{K(k)K(k_1')}$$

$$k_1 = \frac{\sinh(\pi a/2h)}{\sinh(\pi b/2h)}$$

$$\frac{K(k)}{K'(k)} = \begin{cases} \left[\frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k'}}{1+\sqrt{k'}}\right)\right]^{-1}. & \text{for } 0 \le k \le 0.7 \\ \frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k}}{1+\sqrt{k}}\right). & \text{for } 0.7 \le k \le 1 \end{cases}$$



We discussed another form of planar transmission line, which is a coplanar line. So, we can see that, here, the ground plane and the central conductor, they are on the same plane. And this configuration has the advantage that mounting of lumped elements in shunt or series configuration becomes much easier. Now, it supports even or odd quasi TEM modes, depending upon whether the electric fields in these two slots are in the opposite or in the same direction. For this type of line, we can find out the characteristic impedance.

The characteristic impedance is given by Z naught equal to 30 pi divided by root epsilon effective. Epsilon effective can be calculated for this type of geometries into capital K, small k dash divided by capital K small k. Now, this capital K represents the complete elliptic function of the first kind, and capital K dash is actually its complementary function. The small k is defined as a by b where a is S by 2, S is the width of the central strip, b is S by 2 plus W, so W is the width of the slot. And therefore k will become S by 2 divided by S by 2 plus W. And K dash is 1 minus k square. So, we see that the characteristic impedance is going to be a function of a geometry W, S etcetera, and also the dielectric constant epsilon r.

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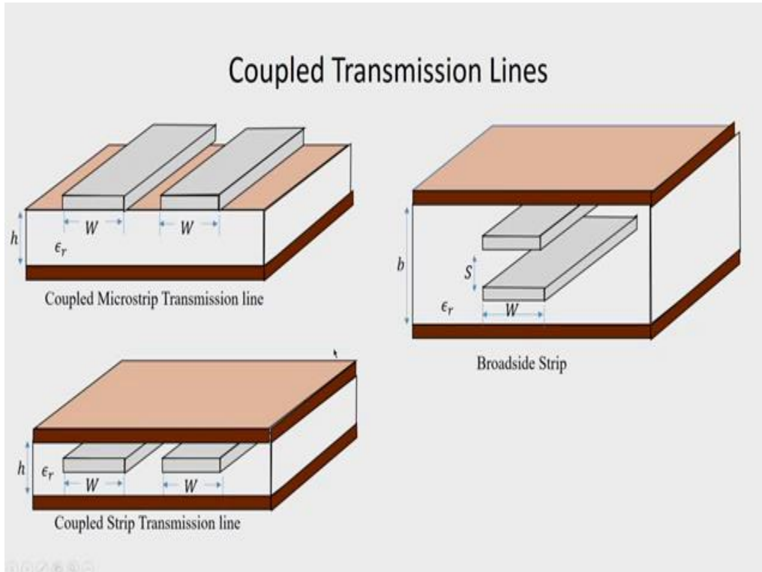

We discussed another interesting transmission line, which is a coupled transmission line. In coupled transmission lines, we have coupling between the conductors, which are either spaced side by side or one above the other. And this type of line, we can have in a microstrip from or in the stripline form. In microstrip form, we have the ground plane, the dielectric material, and the 2 conductors on the dielectric material. Whereas, in coupled strip transmission line we have another ground plane above, and these strips may be side by side as shown in the figure, or they may be one above the other or the broad side strip geometry. So, we have different configurations of coupled strip lines.

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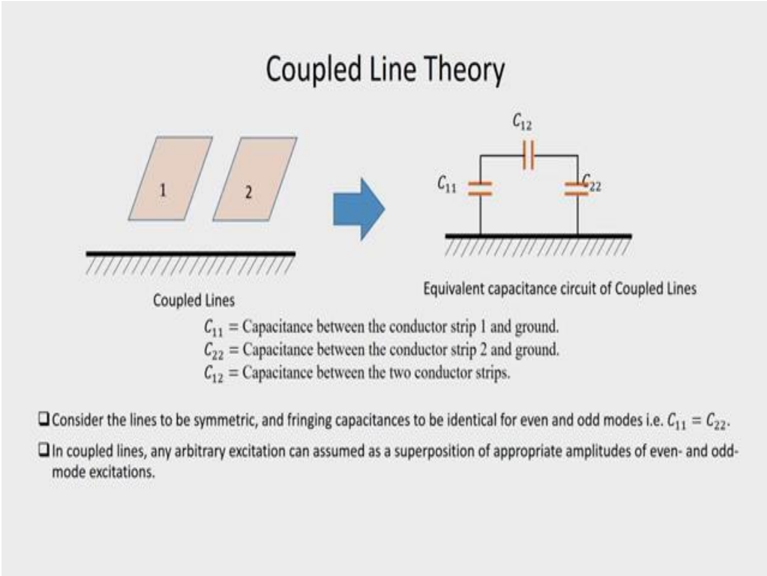

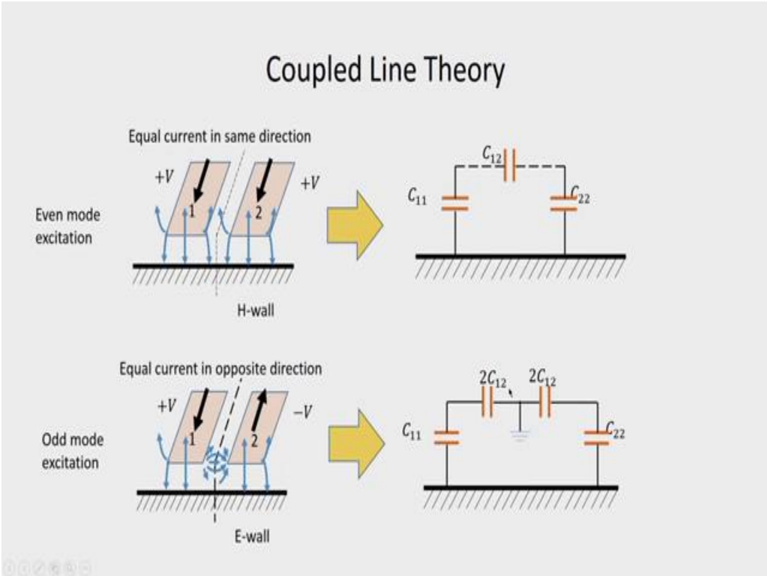

Now, this type of coupled lines can be theoretically analyzed considering the models as shown. So, here we have 2 conductors marked 1 and 2, and then we have the substrate and then the ground. So, we can get an equivalent capacitance model for this type of conductor geometries, as shown in the figure. We have C11 for the conductor 1, C22 for the conductor 2, and C12 is the capacitance formed between conductor 1 and 2.

Now, let us consider that the lines to be symmetric and fringing capacitance because there will be some fringing capacitance also due to the fringing of the fields. And since we are doing an analytical treatment, we assume that we have the capacitances identical for the even and odd modes that mean we have C11 equal to C22.

In coupled lines, arbitrary excitation can be assumed to be a superposition of appropriate amplitudes of even and odd mode excitation. That means what we do. If we have any arbitrary excitation we express it as superposition of even mode, which means when both the conductors are at the same potential, an odd mode when magnitude of the potential of both the conductors is same, but the sign is different.

So, we have for even mode excitation. Both the conductors are at V, and as a result we will have an open circuit created at the middle or an H-wall and equal current flows in the same direction. Now, this type of a situation can be represented because once you have an open circuit or H-wall here, there will be an open circuit between the 2 conductors and therefore we represent it by this dotted line because C12 has been made open now and C11 and C22 are considered to be identical.

Similarly, when we have odd mode excitation that is conductor 1 is having V, conductor 2 minus V, then we will have equal currents in the opposite direction, and this will lead to an equivalent circuit. Because now this will be an E-wall or effectively a short circuit, and this C12 will be now distributed to both the parts of the circuits as 2C12.

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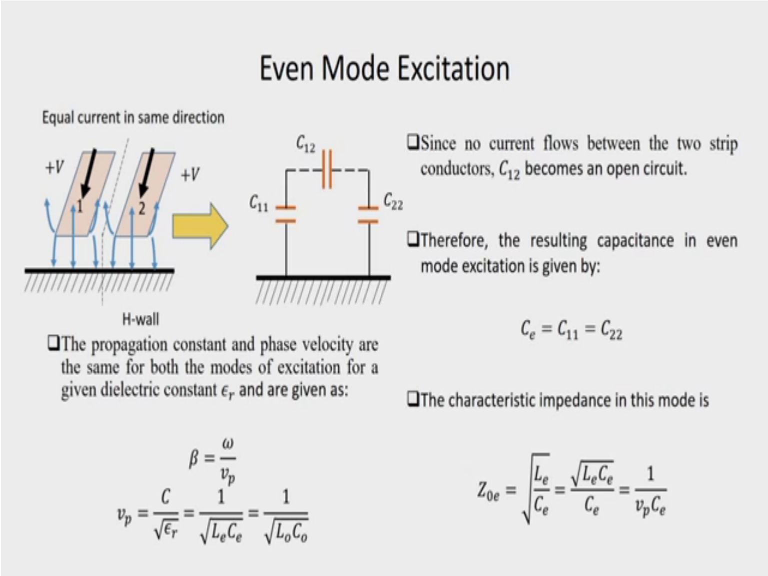
$$\beta = \frac{\omega}{v_p}$$

$$v_p = \frac{C}{\sqrt{\epsilon_r}} = \frac{1}{\sqrt{L_e C_e}} = \frac{1}{\sqrt{L_o C_o}}$$

$$Z_{0e} = \sqrt{\frac{L_e}{C_e}} = \frac{\sqrt{L_e C_e}}{C_e} = \frac{1}{v_p C_e}$$



The propagation constant and phase velocity are the same for both the modes of excitation for a given dielectric constant epsilon r, and these two are given as propagation constant is beta equal to omega by Vp, where Vp is the phase velocity. And Vp is equal to C by root epsilon r, C is the velocity of light in free space, and this can be written as 1 by root Le Ce, this Le and Ce these are the even mode quantities and equal to 1 by root Lo Co odd mode quantities, inductor, and capacitor.

Since no current flows between the two strip conductors, C12 becomes an open circuit now, and therefore the resulting capacitance is even mode excitation. It is given by Ce equal to C11 is equal to C22. And characteristic impedance, Z naught e can be found as 1 by Vp Ce.

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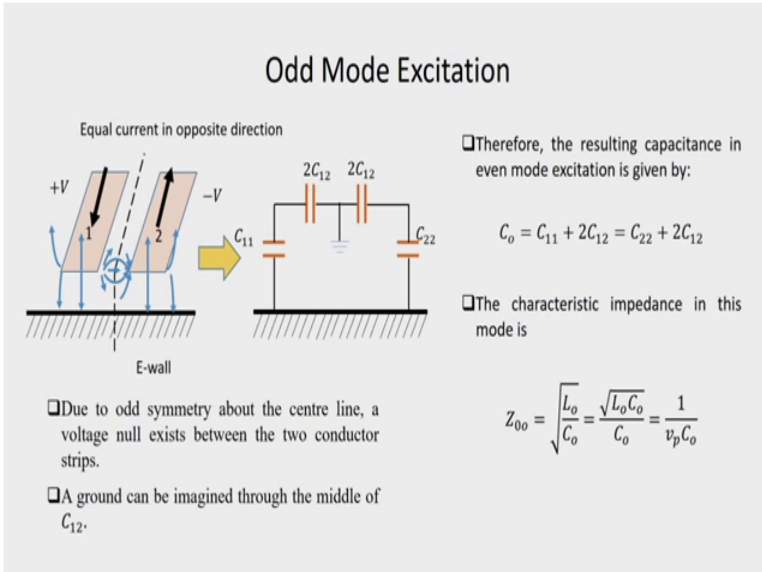
$$C_e = C_{11} + 2C_{12} = C_{22} + 2C_{12}$$

$$Z_{0o} = \sqrt{\frac{L_o}{C_o}} = \frac{\sqrt{L_o C_o}}{C_o} = \frac{1}{v_p C_o}$$



We now consider the odd mode of excitation. As we have already mentioned that for this mode of excitation, an E-wall is formed at the line of symmetry. And we have the equivalent capacitor represented as C_{11} in series with $2C_{12}$ and this C_{12} is now distributed C_{11}, so in odd mode excitation as we have seen, as we have said that an E-wall is formed at the plane of symmetry and therefore we have the equivalent capacitor model as shown here and C_{12} is now split into both the parts of the circuit which is value doubled.

And the resulting total capacitance for the odd mode is C_{11} plus $2C_{12}$, which is equal to C_{22} plus $2C_{12}$. The characteristic impedance in this mode is given by Z naught o, which is equal to 1 by Vp into C_o. So, following this type of analysis we can find out the characteristic impedance for this type of coupled line system. So, that brings to the end of our discussion.

On planar transmission lines, we have discussed various forms of transmission lines, we have talked about stripline, microstrip line, we have discussed slot lines, then coplanar geometries and finally the coupled lines. Next we will discuss the lumped elements that are used in microwave integrated circuit, their characteristics and some of the methodologies which are adopted in their fabrication.