IMPERIAL COLLEGE LONDON

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2009

EEE PART III/IV: MEng, BEng and ACGI

Corrected Copy

MICROWAVE TECHNOLOGY

Wednesday, 6 May 10:00 am

Time allowed: 3:00 hours

There are SIX questions on this paper.

Answer FOUR questions.

All questions carry equal marks

Any special instructions for invigilators and information for candidates are on page 1.

Examiners responsible	First Marker(s) :	S. Lucyszyn
	Second Marker(s) :	C. Papavassiliou

Special instructions to candidates

Permeability of free space, $\mu_o = 4\pi \times 10^{-7}\ H/m$

Permittivity of free space, $\varepsilon_o \approx 8.854\ pF/m$

The Questions

1. An ideal semiconductor has a dielectric constant of 11.9 when dark. It is then illuminated by a laser beam, having a wavelength with a corresponding energy that is greater than the bandgap energy of the semiconductor, creating a plasma of electron-hole pairs inside the semiconductor. The photoconductivity of the plasma region, σ, is given by the following equation:

$$\sigma = \frac{\sigma_o}{1 + j\left(\frac{\omega}{\omega_\tau}\right)} \qquad (1.1)$$

where σ_o is the bulk photoconductivity at DC; ω is the angular frequency and ω_τ is the carrier collision angular frequency. The bulk photoconductivity at DC and the plasma angular frequency ω_p can be defined by the following equations:

$$\sigma_o = \frac{Ne^2}{m\omega_\tau} \qquad (1.2) \qquad \text{and} \qquad \omega_p = \sqrt{\frac{Ne^2}{m\varepsilon_o}} \qquad (1.3)$$

where N is the carrier density, e is the charge of an electron, m is the effective optical mass of the electron and ε_o is the permittivity of free space.

a) Derive an expression for the effective relative permittivity of the plasma region.

[10]

b) From the equation derived in 1(a), calculate the effective relative permittivity and loss tangent of the plasma region at 300 GHz, given σ_o = 3,300 S/m and carrier collision frequency of 3 THz.

[5]

c) A plane wave propagates through a semiconductor having a homogeneous plasma density. Derive an expression for the effective relative permittivity, given the complex refractive index.

[3]

d) Write down an expression for the time and spatial dependency of the electric field in 1(c), as it propagates in one dimensional space, in terms of the complex refractive index. Define all variables used.

[2]

2.

a) A TEM wave propagates within a lossless dielectric, above a perfectly flat metal conducting sheet, in the $+z$ direction and with its electric field in the $-x$ direction and magnetic field in the $+y$ direction. Give the mathematical relationship between the induced electric field vector at the surface of a normal conductor and its incident tangential magnetic field. Define all variables used.

[2]

b) With the scenario in 2(a), and the help of a diagram, explain the directions of the resulting electric field vector and explain why the wave is not pure TEM just above the surface of the normal conductor.

[5]

c) With the scenario in 2(a), define the pointing vectors in the dielectric medium well above the surface of the normal conductor and within the conductor just below its surface. Define the conduction current density as a function of conductivity and also with respect to its depth inside the metal. Define all variables used.

[5]

d) State the mathematical relationship between the conduction current density, as a function of depth, and surface current density. The time dependence can be ignored. Hence, from first principals with the classical skin depth model, show that the propagation constant acts a constant of proportionality.

[4]

e) Using first principals, prove that the conduction current density leads the surface current density by 45° at microwave frequencies.

[4]

3.

a) With the aid of a diagram, illustrate a microstrip 180° 3dB directional coupler (also know as a *Rat Race* coupler). Indicate the characteristic impedances of the various sections.

[3]

b) Describe 6 key features of the Rat Race coupler.

[6]

c) If a Rat Race coupler is required to be implemented at around 900 MHz, suggest a suitable way to integrate it into a miniature hybrid circuit. In addition, state three main penalties for using this alternative method of implementation.

[4]

d) For a 50 Ω system, calculated the component values for 3(c) and the corresponding cut-off frequency. In addition, explain why this implementation cannot work.

[4]

e) What kind of applications is the Rat Race coupler used for and identify a significant problem if lossy components are employed in its realization?

[3]

4.

a) From first principles, derive the expression for the normalised input impedance to a lossless transmission line of arbitrary length that is terminated with arbitrary load impedance. State any assumptions made.

[8]

b) From the expression derived in 4(a), sketch the normalized input impedance against physical length, for an open circuit load impedance. If the lengths of the transmission line are $\lambda g/8$ and $3\lambda g/8$, calculated the effective lumped-element component values at 2.45 GHz, for 50 Ω transmission lines.

[6]

c)

i) State one advantage of an open circuit stub, when compared to a short circuit stub. Also, give a common practical application for the open circuit stub and state any assumption about its electrical length.

[3]

ii) State one advantage of a short circuit stub, when compared to an open circuit stub. Also, give a common practical application for the short circuit stub and state any assumption about its electrical length.

[3]

5. Consider a metal-pipe rectangular waveguide with internal dimensions shown in Figure 5.1

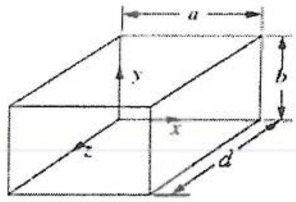


Figure 5.1

a) Give an expression for the ideal (i.e. lossless) TE_{10} guided wavelength for the structure in Figure 5.1, in terms of free-space wavelength and lossless cut-off frequency. Define all variables used. You may assume that the dimensions in Figure 5.1 are $d > \alpha > b$.

[3]

b) Using the resulting equation in 5(a), calculate the ideal cut-off frequency and guided wavelength when $a = 200$ μm and the operating frequency is 918 GHz.

[2]

c) Derive an expression, from first principles, for the ideal (i.e. lossless) TE_{mnl} resonant mode frequencies for the structure given in Figure 5.1

[4]

d) Reduce the equation derived in 5(c) for the TE_{101} mode with $a = 2b = d/\sqrt{2}$. Also, in this specific case, how does the TE_{101} resonant frequency relate to the TE_{10} cut-off frequency?

[2]

e) Using the equations in 5(d), calculate the ideal resonant frequency for a 200 μm wide waveguide.

[2]

f) Define the unloaded quality factor for a general resonator, in terms of energy and power.

[1]

g) Derive a simple expression for the free-space wavelength of a TE_{101} mode, λ_{101_ideal}, in terms of its width, given that $a = 2b = d/\sqrt{2}$.

[2]

h) Calculate the skin depth δ of gold at the ideal resonant frequency found in 5(e), given the value of bulk conductivity of 4.517×10^7 S/m.

[2]

i) Using the results from 5(g) and 5(h), calculated the approximate unloaded Q-factor at resonance, for $a = 2b = d/\sqrt{2} = 200\ \mu m$, given the following equation:

$$Q_U = \frac{\lambda_{101\_ideal}}{4\delta}\left\{\frac{2b\left(a^2+d^2\right)^{\frac{3}{2}}}{\left[2b\left(a^3+d^3\right)+ad\left(a^2+d^2\right)\right]}\right\} \quad (5.1)$$

[2]

6. A simple low frequency microwave FET amplifier can be realised by employing conventional microstrip transmission lines for its interconnects, impedance matching and biasing networks.

a) Describe the main problem with conventional microstrip circuits when trying to ground the source connection of a FET? Give two examples of catastrophic circuit failure. Also, state how a conventional microstrip circuit can be modified to reduce this problem.

[3]

b) Compare and contrast the manufacture of TFMS with conventional microstrip. Use suitable illustrations, showing variables for the main cross-sectional dimensions.

[4]

c) Write simple equations, to a first-order approximation, to justify each of the following claims:

i) the characteristic impedances of a TFMS and conventional microstrip line can be equal (hint: use the variables defined in 6(b)).

[3]

ii) losses in a TFMS are higher than in a corresponding conventional microstrip line.

[4]

d) Figure 6.1 shows the photograph of a MMIC amplifier, employing TFMS transmission lines. By inspection, draw the equivalent circuit model of this amplifier and state the role of the two longest TFMS transmission lines. What kind of amplifier topology is this?

[6]

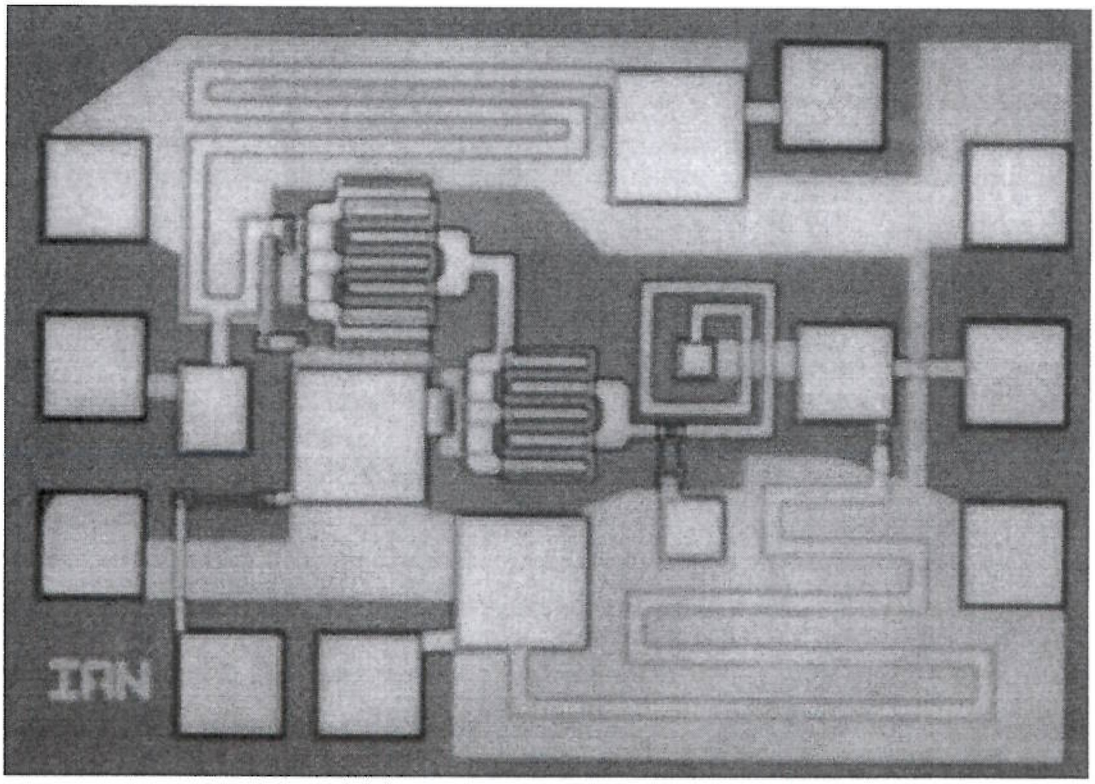


Figure 6.1

The Solutions for E3.18, 2009

Model answer to Q 1(a): New Derivation

$$\sigma = \left(\frac{\omega_p^{\ 2}}{\omega_\tau}\right)\frac{\varepsilon_o}{1+\left(\frac{\omega}{\omega_\tau}\right)}$$

$$\varepsilon_{eff} = \varepsilon_o\varepsilon_r + \frac{\sigma}{j\omega}$$

$$\varepsilon_{reff} = \frac{\varepsilon_{eff}}{\varepsilon_o} = \varepsilon_r - j\left(\frac{1}{\omega}\right)\frac{\omega_p^{\ 2}}{\omega_\tau + j\omega}$$

$$\varepsilon_{reff} = \varepsilon_r - \left(\frac{\omega_p^{\ 2}}{\omega}\right)\left[\frac{\omega + j\omega_\tau}{\omega^2+\omega_\tau^{\ 2}}\right]$$

$$\varepsilon_{reff} = \left\{\varepsilon_r - \left(\frac{\omega_p^{\ 2}}{\omega}\right)\left[\frac{\omega}{\omega^2+\omega_\tau^{\ 2}}\right]\right\} - j\left\{\left(\frac{\omega_p^{\ 2}}{\omega}\right)\left[\frac{\omega_\tau}{\omega^2+\omega_\tau^{\ 2}}\right]\right\}$$

[10]

Model answer to Q 1(b): Computed example

$\varepsilon_r = 11.9 \quad \sigma_o = 3300\,S/m \quad \omega_\tau = 2\pi\times 3\,THz \quad \omega = 2\pi\times 300\,GHz$

$$\sigma_o = \left(\frac{\omega_p^{\ 2}}{\omega_\tau}\right)\varepsilon_o \quad \therefore \omega_p = \sqrt{\frac{\omega_\tau\sigma_o}{\varepsilon_o}} = 265\times 10^{12}$$

$$\therefore \varepsilon_{reff} = -(184 + j1957)$$

$$\tan\delta = \left|\frac{\varepsilon_{reff}''}{\varepsilon_{reff}'}\right| = 10.6$$

[5]

Model answer to Q 1(c): New Derivation

The complex refractive index $n = n' - j\,n''$ is the square root of the complex effective relative permittivity $\varepsilon_{reff} = \varepsilon_{reff}' - j\varepsilon_{reff}''$. Therefore:

$$\varepsilon_{reff} = n^2$$

$$\varepsilon_{reff}' = (n')^2 - (n'')^2$$

$$\varepsilon_{reff}'' = 2n'n''$$

[3]

Model answer to Q 1(d): Derivation

The electric field can be determined at any time t and any position z by $E(t,z) = E(0,0)e^{j\omega t}e^{\gamma z}$, where the propagation constant $\gamma = jkm$ and the modified wavenumber $k_m = n\,k_o$ and the wavenumber in free space $k_o = 2\pi/\lambda_o$ and λ_o is the wavelength in free space.

[2]

Model answer to Q 2(a): Bookwork

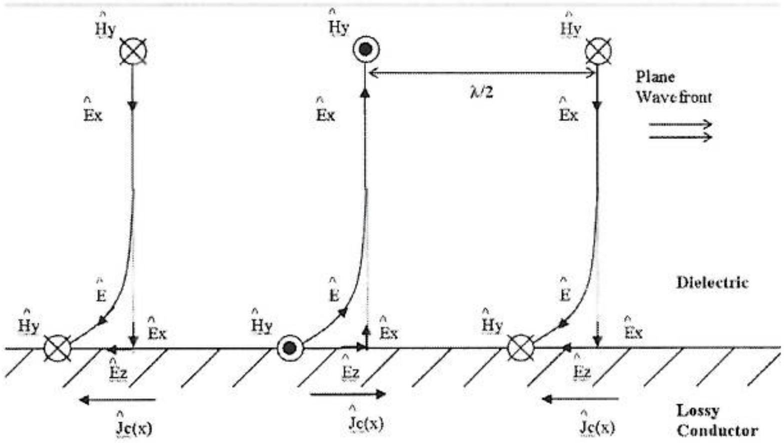
Ez(x=0) = Zs **Js,** where **Js = n** x **Hy** and **n** = normal unit vector and Zs = surface impedance

Therefore, when σo is made finite, a tangential electric field exists, since the surface impedance is no longer zero.

[2]

Model answer to Q 2(b): Bookwork

Since Ez(0) exists, the resultant E-field in the normal metal leans forward, just above the surface of the conductor, i.e. **E** = **x**Ex + **z**Ez. 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.



[5]

Model answer to Q 2(c): Bookwork

Well above the surface the pointing vector **Pz = E-x** x **Hy**.
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) x **Hy**(x), where **Ez**(x) = **Ez**(0)e-γx and Hy = Ez/Zs. If a wave propagates inside the metal, the associated E-field will induce a conduction current, **Jc(x)** = σo **Ez(x)** = **Jc**(0)e-γx.

[5]

Model answer to Q 2(d): Bookwork

$$Js = \int_0^\infty Jc(x).dx$$

$$Js = \frac{Jc(0)}{-\gamma}\left[e^{-\gamma x}\right]_0^\infty$$

$$\therefore Jc(0) = \gamma Js$$

$$\gamma = \sigma_o Zs$$

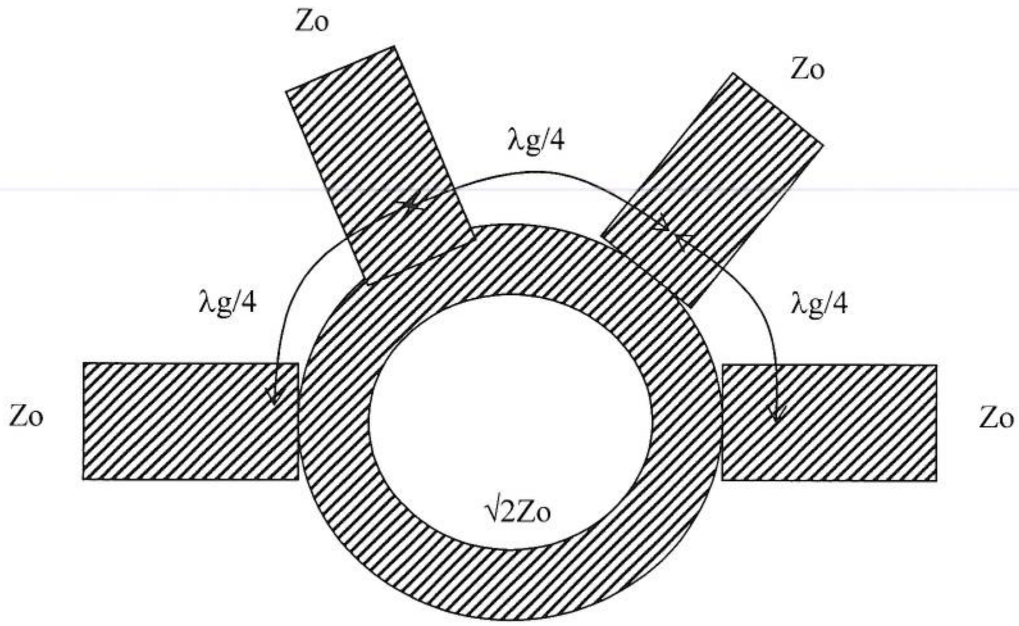
[4]

Model answer to Q 2(e): Bookwork

At the surface of the conductor, the conduction current leads the surface current by 45°, since **Jc(0)** = σo Zs **Js** = σo ($\sqrt{2}\, e^{+j\pi/4}$ Rs) **Js**

[4]

Model answer to Q 3(a): Bookwork



[3]

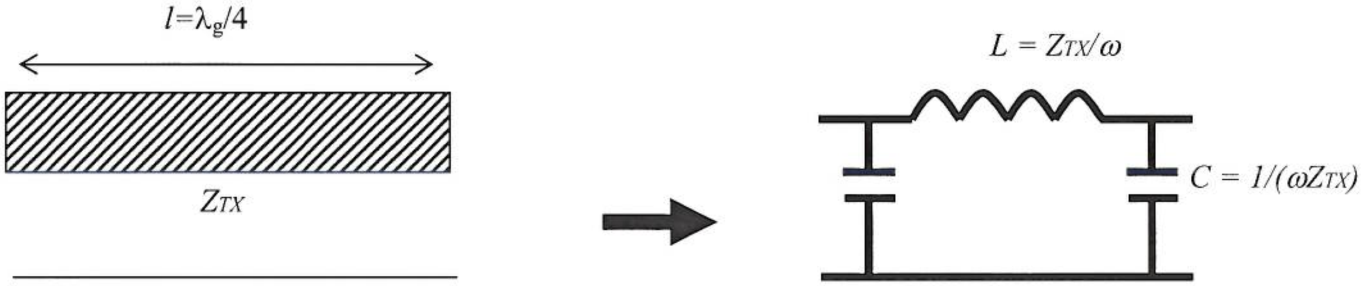
Model answer to Q 3(b): Bookwork

- 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 lower loss and higher power applications
- Simple design but very large
- Meandered lines are possible for lower frequency applications

[6]

Model answer to Q 3(c): Bookwork

The best way to integrate such a low frequency coupler is to employ the lumped-element equivalent circuit for each section of $\lambda_g/4$ transmission line.



Since lumped-element components have a lower Q-factor, when compared to distributed-element components, there is an insertion loss penalty. Also, because this π-network is clearly a low-pass filter, having a cut-off frequency, $f_c = \frac{1}{2\pi\sqrt{LC}}$, there is a bandwidth penalty. Finally, since spiral inductor would be used, the useful bandwidth may also have unwanted resonances due to parasitic capacitances between the windings.

[4]

Model answer to Q 3(d): Computed example

Here, $Z_{TX} = \sqrt{2}Zo = 70.7\Omega$. There will be a ring of 6 inductors, having an identical inductance of $L = Z_{TX} /(2\pi \times 900\ MHz) = 12.5nH$. At each node in the ring there will be a total shunt capacitance down to ground of $C = 2/(Z_{TX} \times 2\pi \times 900MHz) = 5pF$. The individual π-networks for each section of $\lambda_g/4$ transmission line. However, since the total capacitance at each node is twice the normal value, the low-pass cut-off frequency for the ring will be 637 MHz, which is well below the operating frequency.

[4]

Model answer to Q 3(e): Bookwork

The coupler is designed to give an equal output power from two of its ports, but the phase difference of these two output signals is ideally 180°. This coupler is employed in mixers and other applications where anti-phase signals are required. If lossy components are employed, equal amplitude signals will not emerge from the coupler. FOR 1 BONUS MARK, THIS COUPLER CAN ALSO BE USED IN RF SIGNAL PROCESSING APPLICATIONS, WHERE SUM AND DIFFERENCE SIGNALS ARE REQUIRED (E.G. MONOPOLE RADAR ANTENNAS).

[3]

Model answer to Q 4(a): Bookwork Derivation

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

$$\textit{where, } V_\pm (I_\pm)\ \textbf{represents voltage (current) waves at z = 0}$$
$$\textit{and, } \quad e^{\mp\gamma z}\ \textbf{represents wave propagation in the} \pm \textbf{z direction}$$

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

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

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

The impedance looking into a transmission line terminated with load impedance Z_T is given by:

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

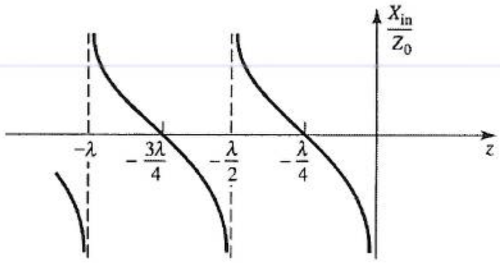
$$\mathbf{Z}in = Zo\frac{\left(Z_T(e^{+\gamma l} + e^{-\gamma l}) + Zo(e^{+\gamma l} - e^{-\gamma l})\right)}{\left(Zo(e^{+\gamma l} + e^{-\gamma l}) + Z_T(e^{+\gamma l} - e^{-\gamma l})\right)}$$

$$\textit{Therefore, } zin = \frac{\mathbf{Z}in}{Zo} = \frac{z_T + \tanh(\gamma l)}{1 + z_T \tanh(\gamma l)} \Rightarrow \frac{z_T + j\tan\theta}{1 + jz_T \tan\theta}\ \textbf{for a lossless line}$$

[8]

Model answer to Q 4(b): Bookwork and Computed Example

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∞ to 0 and 0 to +j∞ as l increases from 0 to λg/4 and λg/4 to λg/2. This is useful for realising any value of "effective" capacitance or inductance over a narrow bandwidth.



If the lengths of the transmission line are λg/8 and 3λg/8, the effective lumped-element component values at 2.45 GHz, for 50 Ω transmission lines are:

$$Zin = -j\frac{50}{\tan(45^o)} = -j50 \equiv \frac{1}{j\omega C} \rightarrow C = 1.3pF$$

$$Zin = -j\frac{50}{\tan(135^o)} = j50 \equiv j\omega L \rightarrow L = 3.25nH$$

[6]

Model answer to Q 4(c): Bookwork

Unlike a short circuit stub, an open circuit stub does not require any through-substrate plated via holes. These can be very difficult and/expensive to implement. One common application is a simple band-stop filter, if the open circuit stub is a quarter-wavelength long, since the short circuit at the input will reflect the incident wave.

[3]

Unlike the open circuit stub, short circuit stubs do not suffer from radiation losses, although they can still excite unwanted substrate modes. One common application is in DC biasing networks, where the RF short circuit at the termination end is transformed into a RF open circuit that will choke off any RF signal.

[3]

Model answer to Q 5(a): Bookwork

The structure has the ideal (i.e. lossless) dominant-mode guided wavelength given by the following textbook expression:

$$\lambda_{g\_ideal} = \frac{\lambda_o}{\sqrt{1-\left(\frac{\lambda_o}{\lambda_c}\right)^2}} = \frac{\lambda_o}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}$$

where, λ_o is the free space wavelength; $\lambda_c = 2a$ is the ideal cut-off wavelength; a is the internal width dimension of the MPRWG; $f_c = c/2a$ is the ideal cut-off frequency for the dominant TE_{10} mode; and c is the speed of light in free space.

[3]

Model answer to Q 5(b): Calculated Example

For a = 200 μm: $f_c = c/(2a) = 750$ GHz and $\lambda_{g\_ideal} = 566.7\,\mu m$

[2]

Model answer to Q 5(c): Bookwork

The corresponding textbook expression for the resonant frequencies for the TE_{mnl} modes in an ideal (i.e. lossless) cavity is give by:

$$\gamma = \hat{x}\gamma_x + \hat{y}\gamma_x + \hat{z}\gamma_x = jk$$

$$k = |\gamma| = \sqrt{k_x^{\,2} + k_y^{\,2} + k_z^{\,2}}$$

$$k_x = \frac{m2\pi}{\lambda_x} \quad \lambda_x = 2a; \quad k_y = \frac{n2\pi}{\lambda_y} \quad \lambda_x = 2b; \quad k_z = \frac{l2\pi}{\lambda_z} \quad \lambda_z = 2a;$$

$$k_{mnl\_ideal} = \frac{2\pi f_{mnl\_ideal}}{c} \quad \therefore f_{mnl\_ideal} = \frac{c}{2\pi}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2}$$

[4]

Model answer to Q 5(d): New Derivation

For the dominant TE_{101} mode this becomes:

$$f_{101\_ideal} = \frac{c}{2\pi}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} = \sqrt{\frac{3}{8}}\left(\frac{c}{a}\right) \text{ when } d = \sqrt{2}a$$

$$f_{101\_ideal} = \sqrt{1.5}\,f_c \text{ when } d = \sqrt{2}a$$

[2]

Model answer to Q 5(e): Computed Example

For a = 200 μm: b = 100 μm, d = 283 μm, $f_{101\_ideal} = 918$ GHz

[2]

Model answer to Q 5(f): Bookwork

$$Q_U = 2\pi\frac{Time - average \;\; energy \;\; stored \;\; at\ a\ resonant \;\; frequency}{Energy \;\; dissipated \;\; in\ one \;\; period \;\; of\ this \;\; frequency} \text{ (Dimensionless)}$$

$$Q_{Uo} = \omega_{oo}\frac{U}{P_{DISS}} \text{ (Dimensionless)}$$

[1]

Model answer to Q 5(g): New Derivation

$$\lambda_{101\_ideal} \equiv \frac{2\pi c}{\omega_{101\_ideal}} = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{l}{d}\right)^2}} = 2\sqrt{\frac{2}{3}}a \quad for \quad d = \sqrt{2}a$$

[2]

Model answer to Q 5(h): Calculated Example

$$\delta = \sqrt{\frac{2}{\omega_{101\_ideal}\mu_o\sigma_o}} = 78nm \quad at\ 918\ GHz$$

[2]

Model answer to Q 5(i): Calculated Example

$$Q_U = \frac{\lambda_{101\_ideal}}{4\delta}\left\{\frac{2b\left(a^2+d^2\right)^{\frac{3}{2}}}{\left[2b\left(a^3+d^3\right)+ad\left(a^2+d^2\right)\right]}\right\} = \frac{2}{\delta_{So}}\left(\frac{3\sqrt{2}a}{4(5\sqrt{2}+1)}\right) \quad for \quad d = \sqrt{2}a$$

$$\therefore Q \approx 674 \quad for\ a = 200\mu m\ and\ d = \sqrt{2}a$$

[2]

Model answer to Q 6(a): Bookwork

The main problem with conventional microstrip circuits, when trying to ground the source connection of a FET, is that the through-substrate metal-plated via connection has both parasitic inductance and resistance. This can prevent an oscillator from oscillating or cause amplifiers to oscillate. Both these examples represent catastrophic circuit failure.

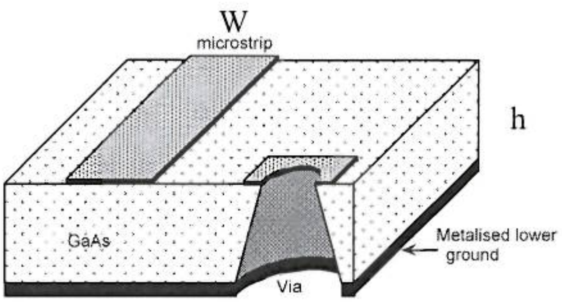
A conventional microstrip circuit can be modified to reduce this problem by turning it into thin-film microstrip.

[3]

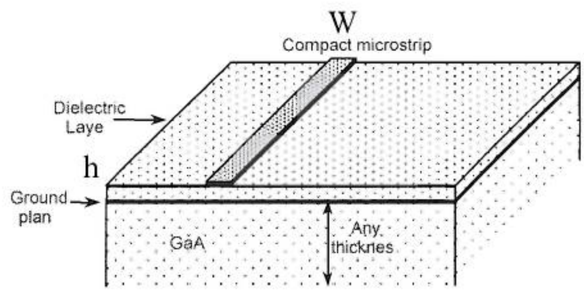
Model answer to Q 6(b): Bookwork

When compared to conventional microstrip, the ground plane is brought up to the top of the substrate. A thin layer of non-conductor defines the dielectric that separates the ground plane from the main signal line. The dielectric layer can be much thinner for TFMS than with conventional microstrip, and easily deposited using a spin-on or lamination bonding process

Conventional Microstrip



Thin-Film Microstrip (TFMS)



[4]

Model answer to Q 6(c): Bookwork

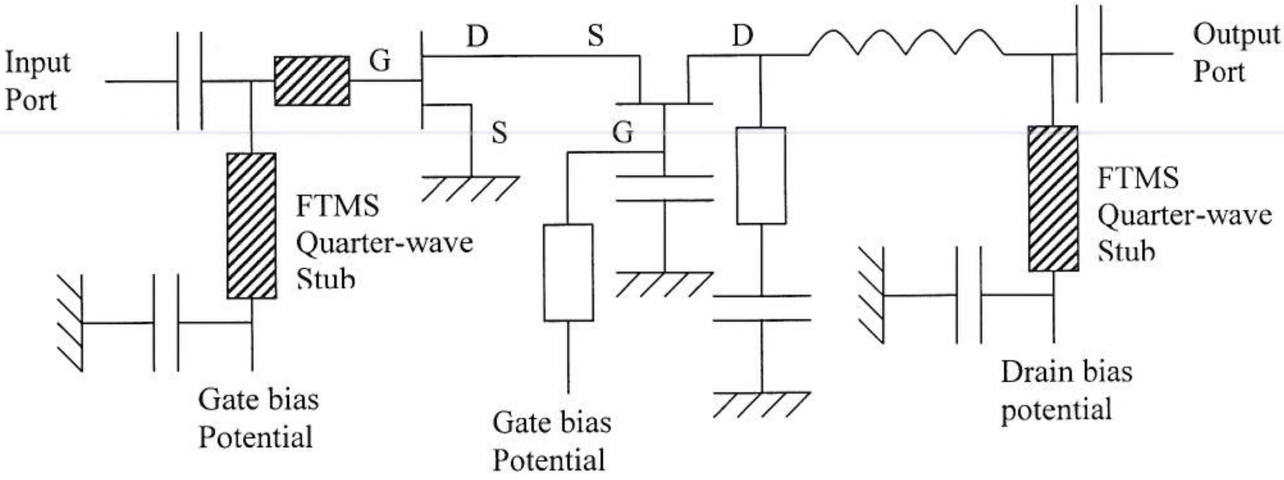
i) The characteristic impedances Zo of a TFMS and conventional microstrip line can be the same (hint: use the variables defined in 6(b)). It will be found that $Zo \propto \frac{W_{effective}}{h}$. Therefore, as long as this ratio stays the same then characteristic impedance can stay the same. This means that W will be much narrower with TFMS than with conventional microstrip.

[3]

ii) losses in a TFMS is higher than a corresponding conventional microstrip line. The reason is that power dissipated as heat is defined by: $P_{DISSIPATED} = |Js|^2 Rs$, where Js is the surface impedance and Rs is the surface resistance. Since Js is directly proportional to the conduction current density $Jc(0)$ [A/m^2], it follows that as the cross-sectional area of the microstrip's signal track is much smaller with TFMS then $Jc(0)$, Js and $P_{DISSIPATED}$ are all much greater.

[4]

Model answer to Q 6(d): Bookwork



The two largest TFMS transmission lines are used as quarter-wave stubs, which act as DC bias chokes. The amplifier has a cascode topology.

[6]