Paper Number(s): **E4.18**
AM5

IMPERIAL COLLEGE OF SCIENCE, TECHNOLOGY AND MEDICINE
UNIVERSITY OF LONDON

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2002

MSc and EEE PART IV: M.Eng. and ACGI

RADIO FREQUENCY ELECTRONICS

Thursday, 2 May 10:00 am

There are SIX questions on this paper.

Answer FOUR questions.

Corrected Copy

Time allowed: 3:00 hours

Examiners responsible:

First Marker(s): Lucyszyn,S

Second Marker(s): Papavassiliou,C.

Special instructions for invigilators:	This is a Closed Book examination. A Smith Chart is to be distributed. Filter Curves and Tables are to be distributed.
Information for candidates:	This examination is Closed Book. A Smith Chart is provided and, if used, you must attach this to your answer book. Filter Curves and Tables are provided.

1(a) From first principles, derive an expression for both propagation constant and intrinsic impedance, in terms of the loss tangent of a homogeneous isotropic medium. Clearly define all the variables used.

[8]

(b) Using the expressions derived in 1(a), show that within a good conductor the attenuation constant is equal to the phase constant and also that the surface resistance is equal to the surface reactance.

[7]

(c) A printed circuit board (PCB) has 17 [μm] thick copper layers, having a bulk DC conductivity of 5.8 x 10^7 [S/m]. With a surface current density of 0.5 [A/m], calculate the following parameters for a frequency 900 [MHz]:

(i) Skin depth
(ii) Wavelength
(iii) Surface conduction current density
(iv) Surface power dissipation.

[5]

(d) For the PCB and frequency in 1(c), calculate the isolation offered by a single copper layer.

[5]

2. A 500 [MHz] small-signal amplifier has an output impedance of 20 – j 15 [Ω]. Using the Smith charts provided, design suitable impedance matching networks to transform this impedance to 50 [Ω]:

(i) using a quarter-wavelength transformer

[5]

(ii) using a short-circuit stub

[10]

(iii) using a discrete inductor and capacitor

[5]

(iv) make general comments about the loaded-Q factor of the matching networks in (i), (ii) and (iii) and how this relates to the resulting bandwidth of the networks.

[5]

3(a) With the use of simple illustrations for the attenuation against frequency curves, describe the differences between Butterworth, Chebyshev and Elliptical-function filters. Also, comment on the group delay characteristics for these filters.

[5]

(b) Given prototype low-pass filter attenuation curves and tables for the corresponding normalised element values (see Filter Curves and Tables provided), design an *L-C* lumped-element band-pass filter that meets the following specifications:

Centre Frequency, f_O	500 [MHz]
3 dB Bandwidth, B	50 [MHz]
Attenuation Bandwidth	100 [MHz]
Pass-Band Ripple (Peak-to-Peak)	0.1 [dB]
Stop-Band Attenuation	45 [dB]
Input Impedance, R_{IN}	100 [Ω]
Output Impedance, R_{OUT}	50 [Ω].

[20]

4(a) With the aid of a diagram, describe the S-parameter representation of a linear two-port circuit, stating the precise definitions of all parameters and the main power specifications.

[8]

(b) State which RF components best describe the following S-parameter matrices and calculate any relevant power specifications:

(i) $$[S]=\begin{pmatrix} 0 & e^{-j720} \\ e^{-j720} & 0 \end{pmatrix}$$

(ii) $$[S]=\begin{pmatrix} 0 & 0.07e^{-j30} \\ e^{-j60} & 0 \end{pmatrix}$$

(iii) $$[S]=\begin{pmatrix} 0.1e^{-j30} & 0.3e^{-j80} \\ 9.7e^{-j80} & 0.15e^{-j60} \end{pmatrix}.$$

[5]

(c) Derive an algebraic expression for the overall S_{11}, of a linear two-port network that is terminated at its output port with a one-port network represented by S_{11L}.

[7]

(d) Referring to the result in 4(c), state the condition for stability for the overall one-port network, for any value of generator source impedance. If the two-port network in 4(b)(ii) is terminated with a load impedance having S_{11L} = 0.5, determine if the overall one-port network is stable.

[5]

5(a) For a transistor, define the transition frequency, f_T, unity gain cross-over frequency, f_S, and maximum frequency of oscillation, f_{MAX}. Explain why a high-power FET amplifier is difficult to design at microwave frequencies when using only a single transistor stage. [5]

(b) A FET has a small-signal equivalent circuit model with corresponding elements having the following values:

$g_{mo} = 15$ [ms]
$C_{gs} = 81$ [fF]
$R_{ds} = 535$ [Ω]
$R_g = 1.13$ [Ω]
$R_i = 12.4$ [Ω]
$R_s = 5.3$ [Ω].

All variables have their usual meaning

Calculate the extrinsic transconductance, transition frequency and maximum frequency of oscillation. Comment on the suitability of this transistor for use in an amplifier for applications in vehicular radars operating at 76.5 [GHz]. [7]

(c) Draw the circuit diagram of a simple class-A amplifier and describe clearly the function of each component. Describe the performance characteristics of a class-A RF power amplifier and state its applications. [5]

(d) Consider a 4 x 75 μm FET with the following specifications:

$V_{gd|BD} = 12$ [V]
$V_p = -1.5$ [V]
$V_k = 0.5$ [V]
$I_{dss} = 60$ [mA]
Small-Signal Power Gain = 4 [dB].

All variables have their usual meaning

Calculate the following for delivering the maximum linear output power:

(i) Optimal bias voltages
(ii) Optimal load impedance
(iii) Values of maximum linear output power and peak output power
(iv) DC power and power dissipated per unit gate width
(v) Drain efficiency and Power-added efficiency. [8]

6(a) With the aid of a diagram and simple S-parameter analysis, explain how a lossless single-stage reflection-topology works. [10]

(b) Explain how an ideal reflection-type attenuator can be implemented and extend the analysis given in 6(a) for this application. [5]

(c) Explain how an ideal reflection-type phase shifter can be implemented and extend the analysis given in 6(a) for this application. [5]

(d) Describe the inherent drawbacks of the applications given in 6(b) and 6(c) at microwave frequencies when non-ideal components are used. How can a useful vector modulator be implemented with non-ideal components. [5]

Model Answers and Mark Schemes — First Examiner: S. LUCYSZYN

Paper Code: E4.18 — Second Examiner: C. PAPAVASSILIOU

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SOLUTIONS

1(a) 1(a) PROPAGATION CONSTANT, $\gamma = jk \equiv \alpha + j\beta$

$$k = 2\pi k = \frac{2\pi}{\lambda}$$

$$\lambda = \frac{v_p}{f}$$

$$v_p = \frac{1}{\sqrt{\mu\varepsilon}}$$

$$\therefore \gamma = j\omega\sqrt{\mu\varepsilon_0}$$

$$\varepsilon_r = \frac{\varepsilon}{\varepsilon_0} = \varepsilon_r' - j\varepsilon_r''$$

$$\tan\delta = \frac{\varepsilon_r''}{\varepsilon_r'} = \frac{\varepsilon''}{\varepsilon'} \quad \text{AND} \quad \sigma = \omega\varepsilon''$$

$$\therefore \gamma = j\omega\sqrt{\mu(\varepsilon' - j\tfrac{\sigma}{\omega})}$$

$$\boxed{\gamma = j\omega\sqrt{\mu\varepsilon'(1-\tan\delta)}}$$

[15] 15

INTRINSIC IMPEDANCE, $\eta = \sqrt{\frac{\mu}{\varepsilon}}$

$$\eta = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon'}} = \sqrt{\frac{\mu/\varepsilon'}{1 - j\sigma/\omega\varepsilon'}}$$

$$\boxed{\eta = \sqrt{\frac{\mu/\varepsilon'}{1 - j\tan\delta}}}$$

[15] 15

1(b) 1(b) $\tan\delta \gg$ For Good conductor $\therefore \sqrt{1 - j\tan\delta} \simeq \sqrt{\frac{\tan\delta}{2}}(1-j)$

$$\therefore \gamma \simeq j\omega\sqrt{\frac{\omega\mu\sigma}{2}}(1+j) \equiv \alpha + j\beta$$

$$\therefore \boxed{\alpha \simeq \beta}$$

[15] 15

$$\eta \simeq \sqrt{\frac{\mu\omega}{2\sigma}}(1+j) \equiv Z_s = R_s + jX_s$$

$$\therefore \boxed{R_s = X_s}$$

[15] 15

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1(c) c)(i) $R_S = 7.827\ [m\Omega]$

$\therefore \delta_o = \dfrac{1}{R_S \sigma_o} = 2.2\ [\mu m]$ [5] 5

(ii) $\lambda = 2\pi\delta_o = 13.8\ [\mu m]$ [5] 5

(iii) $J_c(0) = \dfrac{J_S}{\delta_o}(1+j) = 0.227(1+j)\ [MA/m^2]$ [5] 5

(iv) $P_D(0) \equiv P_S = |J_S|^2 R_S = 2.0\ mW/m^2$ [5] 5

1(d) 1 (d) $\alpha = \dfrac{1}{\delta_o} = 454.5 \times 10^3\ [Np/m]$

$\text{ISOLATION} \equiv \text{PROPAGATION LOSS} = 10\log\left|\dfrac{E(T)}{E(0)}\right|^2$

$\text{ISOLATION} = 10\log e^{-2\alpha T} = -20\alpha T \log\{e\}$

$\text{ISOLATION} = -8.686\,\alpha T\ [dB]$

$\text{ISOLATION} = -67\ [dB]$ [20] 20

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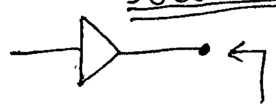
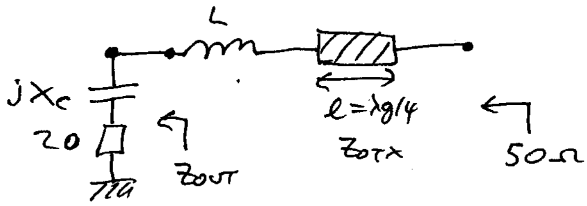
SOLUTIONS (several different solutions acceptable, this is one. C.P.

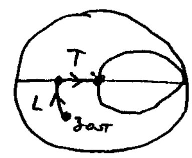
2(a) a)



$Z_0 = 50\Omega$ $Z_{OUT} = 20 - j15\Omega$

$z_{OUT} = \frac{Z_{OUT}}{Z_0} = 0.4 - j0.3$





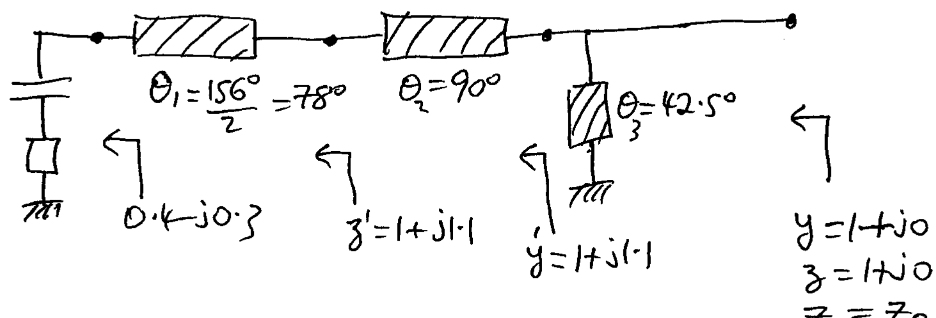
$X_C = -15\Omega$ $\therefore X_L = +15\Omega$

$\therefore L = \frac{15}{\omega} = 4.78\,nH$

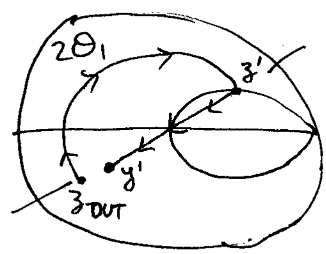
$Z_{OTX} = \sqrt{R_{OUT}\,50} = 31.6\Omega$

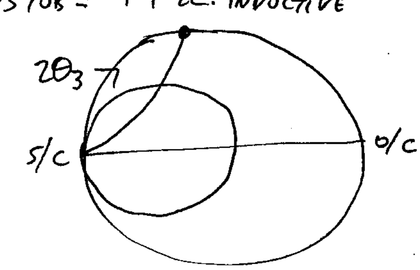
[20] 20

2(b) 2(b)



$b_{STUB} = -1.1$ ie. INDUCTIVE





[40] 40

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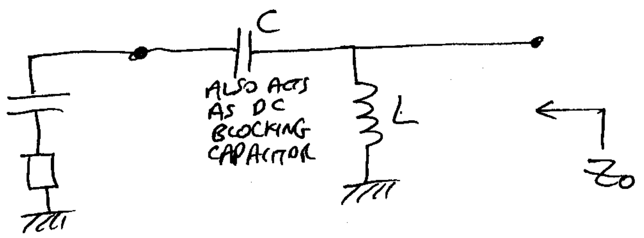
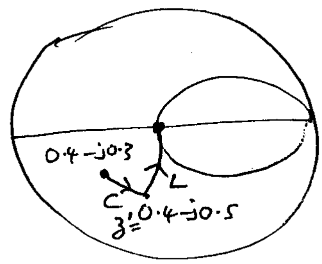
2(c) (c)





$x_C = -0.2 \quad \therefore \quad X_c = -x_c Z_0 = -10\Omega$

$X_c = -\frac{1}{\omega C} \quad \therefore \quad C = 31.8\,pF$

$z' = 0.4 - j0.5$

$\therefore \; y' = \frac{1}{z'} \simeq 1 + j1.22$

$\therefore \; b_L = -1.22$

$B_L = \frac{b_L}{Z_0} = -\frac{1}{\omega L} = -0.0244$

$\therefore \; L = 13\,nH$

[20] 20

2(d) 2(d) ON SMITH CHART LOADED-Q IS ZERO AT THE CENTRE AND INFINITE AT THE EDGE. THIS MEANS THAT THE L-C NETWORK HAS A POINT CLOSER TO THE EDGE OF THE CHART AND, THEREFORE, LOAD-Q IS HIGHER THAN $\lambda g/4$ TRANSFORMER OR SINGLE STUB. ALSO BANDWIDTH WILL BE LESS, SINCE LOADED-Q = $\frac{\text{CENTRE FREQUENCY}}{\text{BANDWIDTH}}$.

SINCE THE STUB SOLUTION HAS MORE LENGTH OF TRANSMISSION LINES, IT WILL OPERATE OVER A NARROW BANDWIDTH COMPARED TO ~~λ/4~~ TRANSFORMER. [20] 20

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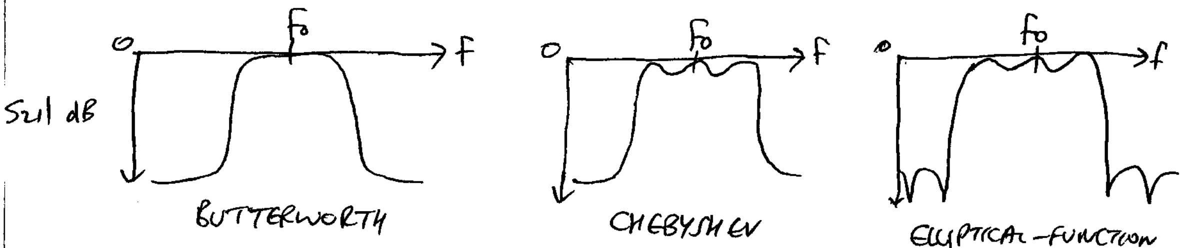
SOLUTIONS

3(a) , (a) BUTTERWORTH FILTERS HAVE A MAXIMALLY FLAT FREQUENCY RESPONSE IN THE PASSBAND. CHEBYSHEV FILTERS EXHIBIT RIPPLES IN THE PASSBAND; THE NUMBER DEPENDING ON THE FILTER'S ORDER. NEITHER THE BUTTERWORTH OR CHEBYSHEV FILTERS HAVE RIPPLES IN THE STOPBAND. ELLIPTICAL-FUNCTION FILTERS HAVE THE SHARPEST ROLL-OFF OF ALL THREE TYPES, WITH THE CHEBYSHEV HAVING A SHARPER ROLLOFF THAN THE BUTTERWORTH. THE ELLIPTICAL-FUNCTION FILTER EXHIBITS RIPPLES IN THE STOPBAND (ZEROS) AS WELL AS RIPPLES IN THE PASSBAND (POLES). CHEBYSHEV AND ELLIPTICAL-FUNCTION FILTERS HAVE A HIGHER GROUP DELAY THAN THE BUTTERWORTH, PARTICULARLY NEAR THE 3dB CUT-OFF FREQUENCY AND SHOULD BE AVOIDED IN APPLICATIONS WHERE PULSE DISTORTION IS CRITICAL.



[20] 20

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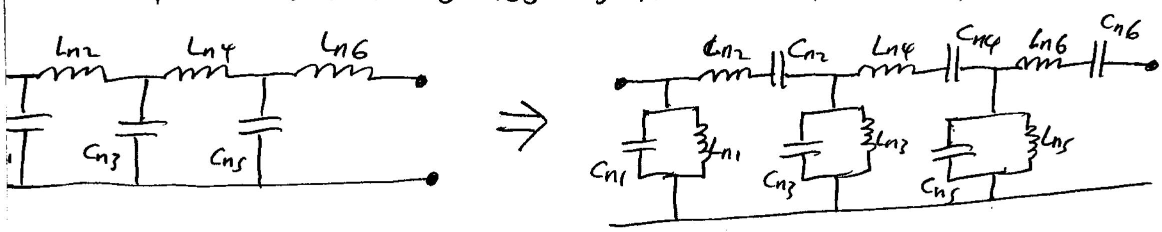
3(b) i) DETERMINE $\frac{F}{F_C} = \frac{B}{BW_{STOP}} = 2$

FROM CURVES: * BUTTERWORTH WITH 0.00 dB RIPPLE NEED >7th ORDER
* CHEBYSHEV WITH 0.01 dB RIPPLES NEED 6th ORDER FOR 46dB REJECTION HAVING $F/F_C = 2$
* CHEBYSHEV WITH 0.10 dB RIPPLES NEED 6th ORDER FOR 50 dB REJECTION HAVING $F/F_C = 2$
THIS IS BEST, AS IT HAS A GOOD OUT-OF-BAND ATTENUATION MARGIN

FOR $\frac{R_{IN}}{R_{OUT}} = \frac{R_S}{R_L} = 2$ THE NORMALISED PROTOTYPE LOW PASS FILTER COEFFICIENTS ARE:

C_{n1}	L_{n2}	C_{n3}	L_{n4}	C_{n5}	L_{n6}
0.414	3.068	0.958	3.712	0.979	2.794



TO DENORMALISE SHUNT CAPACITORS AND INDUCTORS:

$$C_p = \frac{C_n}{2\pi B R} \quad \text{AND} \quad L_p = \frac{1}{C_p \omega_0^2} = \frac{RB}{2\pi f_0^2 C_n}$$

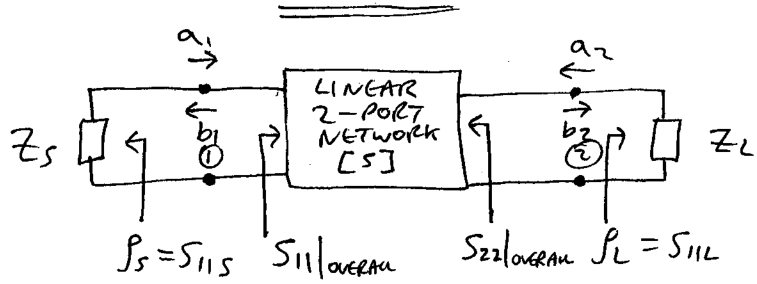
TO DENORMALISE SERIES INDUCTORS AND CAPACITORS:

$$L_s = \frac{L_n R}{2\pi B} \quad \text{AND} \quad C_s = \frac{1}{L_s \omega_0^2} = \frac{B}{2\pi f_0^2 L_n R}$$

$C_{p1} = 26.36\,pF$ $L_{p1} = 3.84\,nH$
$C_{s2} = 0.21\,pF$ $L_{s2} = 488.29\,nH$
$C_{p3} = 61.00\,pF$ $L_{p3} = 1.66\,nH$
$C_{s4} = 0.17\,pF$ $L_{s4} = 590.78\,nH$
$C_{p5} = 62.33\,pF$ $L_{p5} = 1.63\,nH$
$C_{s6} = 0.23\,pF$ $L_{s6} = 444.68\,nH$ [80] 80

SOLUTIONS

4(a) (a)



INCIDENT VOLTAGE WAVE AT ①, $a_1 = \frac{V_{i1}}{\sqrt{Z_{01}}}$

REFLECTED VOLTAGE WAVE AT ①, $b_1 = \frac{V_{r1}}{\sqrt{Z_{01}}}$

INCIDENT VOLTAGE WAVE AT ②, $a_2 = \frac{V_{i2}}{\sqrt{Z_{02}}}$

REFLECTED VOLTAGE WAVE AT ②, $b_2 = \frac{V_{r2}}{\sqrt{Z_{02}}}$

INCIDENT POWER AT ①, $|a_1|^2 = \frac{|V_{i1}|^2}{Z_{01}}$

REFLECTED POWER AT ①, $|b_1|^2 = \frac{|V_{r1}|^2}{Z_{01}}$

$$b_1 = S_{11} a_1 + S_{12} a_2$$
$$b_2 = S_{21} a_1 + S_{22} a_2$$

WHERE $S_{11(22)}$ = VOLTAGE REFLECTION COEFFICIENT AT PORT ① (②)

$S_{21(12)}$ = VOLTAGE TRANSMISSION COEFFICIENT IN THE FORWARD (REVERSE) DIRECTION

MATCHED TRANSDUCER POWER GAIN, $G_{TM} = 10 \log |S_{21}|^2$ [dB]

RETURN LOSS, $RL = 10 \log |S_{11}|^2$ [dB]

[30] 30

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4(b) (b) (i) THIS IS A LOSSLESS TRANSMISSION LINE OF LENGTH $2\lambda_g$. PERFECT MATCHING

(ii) THIS IS AN ISOLATOR. THE INSERTION LOSS IN THE FORWARD DIRECTION IS 0dB AND 23.1 dB IN THE REVERSE DIRECTION. PERFECT MATCHING.

(iii) THIS IS AN AMPLIFIER WITH FORWARD POWER GAIN OF 19.7dB, REVERSE TRANSMISSION LOSS OF 10.45 dB.

IF THE OUTPUT IS MATCHED, THE INPUT RETURN LOSS IS 20dB. IF THE INPUT IS MATCHED, THE OUTPUT RETURN LOSS IS 16.5 dB.

[20] 20

4(c) 4 (c)

$$b_1 = S_{11} a_1 + S_{12} (S_{11L} b_2)$$
$$b_2 = S_{21} a_1 + S_{22} (S_{11L} b_2)$$

$$\therefore b_2 (1 - S_{22} S_{11L}) = S_{21} a_1$$

$$\therefore b_1 = S_{11} a_1 + \frac{S_{12} S_{21} a_1}{(1 - S_{22} S_{11L})}$$

$$\therefore S_{11}\Big|_{OVERALL} = \frac{b_1}{a_1} = S_{11} + \frac{S_{12} S_{21}}{(1 - S_{22} S_{11L})}$$

[30] 30

4(d) 4 (d) IF $|S_{11}|_{OVERALL} \leq 1$ THE CIRCUIT WILL BE UNCONDITIONALLY STABLE. SINCE $S_{22} = 0 = S_{11}$

$$S_{11}\Big|_{OVERALL} = S_{12} S_{21} = 0.07\, e^{-j\pi/2}$$

$$\therefore |S_{11}|_{OVERALL} = 0.07 \ll 1 \therefore \text{UNCONDITIONALLY STABLE}$$

[20] 20

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SOLUTIONS

5(a) 5(a) The transition (or cut-off) frequency, f_T, of a transistor is defined as the frequency at which the short circuit current gain, h_{21}, falls to unity. f_T is also known as the gain-bandwidth product and it is the key figure-of-merit for digital applications.

The unity gain cross-over frequency, fs, of a transistor is defined as the frequency at which the insertion power gain (i.e. ratio of power delivered to the load after the transistor has been inserted to the power delivered to the load before the transistor was inserted), $|S_{21}|^2_{Zo}$, falls to unity. fs > f_T and it can be read directly off the frequency response measurements from a network analyser.

The maximum frequency of oscillation, f_{max}, of a transistor is defined as the frequency at which the maximum (i.e. when both the input & output ports are conjugate matched for maximum power transfer) unilateral (i.e. S_{12}=0) transducer power gain (i.e. ratio of power delivered to the load to the power available from the source), G_{TUmax}, falls to unity. f_{max} > fs and it is the key figure-of-merit for analogue applications.
Now, it can be shown that:

$$f_T = \frac{gmo}{2\pi\, Cgs}$$

Single-transistor high power amplifiers have a large value of Cgs and, therefore, a low value of f_T. As a result, it becomes harder to design the amplifier at higher microwave frequencies because the power gain decreases with increasing frequency at 6 dB/octave.

[20] 20

5(b) 5(b) Extrinsic transconductance is given by the following:

$$gm \sim \frac{gmo}{1 + gmoRs} = 14\,[ms]$$

$$f_T = \frac{gmo}{2\pi Cgs} = 29.5 \text{ GHz}$$

$$f_{MAX} = \frac{f_T}{2}\sqrt{\frac{Rds}{Rg + Ri + Rs}} = 78.6 \text{ GHz}$$

Theoretically, it is just possible to design an amplifier to have some power gain at 76.5 GHz. However, in practice, once the losses in the input and output impedance matching networks are taken into consideration the overall gain of the complete circuit will be negative.

[30] 30

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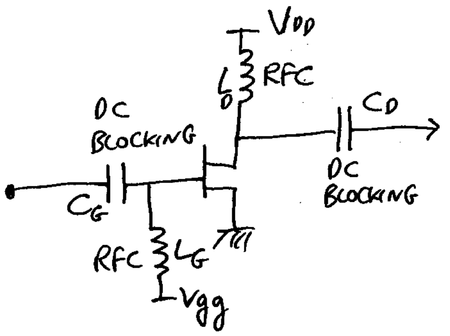
5(c) TOPOLOGY OF CLASS-A AMPLIFIER (NO IMPEDANCE MATCHING STAGES HAVE BEEN SHOWN).



RFC = RADIO FREQUENCY CHOKE

$L_G, L_D \to \infty$

DC BLOCKING CAPACITOR HAS

$C_G, C_D \to \infty$

IN PRACTICE, ALL BIAS COMPONENTS WILL BE NON-IDEAL (i.e. FINITE IN VALUE AND LOSSY).

CLASS-A AMPLIFIERS HAVE THE FOLLOWING CHARACTERISTICS:

* OUTPUT CURRENT FLOWS FOR FULL PERIOD OF INPUT VOLTAGE CYCLE
* NO HIGH ORDER HARMONICS ARE GENERATED AND, THEREFORE, MINIMUM OUTPUT SIGNAL DISTORTION (IN THE TIME-DOMAIN).
* BEST $P_{OUT} - P_{IN}$ LINEARITY
* MAXIMUM DYNAMIC RANGE
* IDEAL FOR NON-CW APPLICATIONS (e.g. AM AND MULTI-CARRIER)
* CONTINUOUS POWER DISSIPATED AS HEAT
* LOW EFFICIENCY (< 50%) AND, THEREFOR, GETS HOT!!

[20] 20

5(d) 5(d)(i) $V_{ds}|_{MAX} = V_{gd}|_{BO} - |V_p| = 10.5V$

$\therefore V_{dd} = \frac{V_{ds}|_{MAX} + V_k}{2} = 5.5\ V$

$V_{gg} = \frac{-|V_p|}{2} = -0.75V$

(ii) $R_L = \frac{V_{ds}|_{MAX} - V_k}{I_{dss}} = 167\Omega$

(iii) $P_{OUT}|_{MAXLIN} = \frac{\left[\frac{V_{ds}|_{MAX} - V_k}{2\sqrt{2}}\right]^2}{R_L} = 75mW \equiv +19dB_m$

$P_{OUT}|_{PEAK} = 2\ P_{OUT}|_{MAXLIN} = 150mW \equiv +23dB_m$

(iv) $P_{DC} = I_{ds}|_Q V_{ds}|_Q = \frac{I_{dss}}{2} \cdot V_{dd} = 165mW$

$P_{IN} = \frac{P_{OUT}}{G} = \frac{75mW}{2.51} = 30mW \equiv +14.75dB_m$

$\therefore P_{DISS} = (P_{IN} + P_{DC}) - P_{OUT} = 120mW$ BUT $W_g = 4 \times 75 = 300\mu m$

$\therefore P_{DISS}/W_g = 0.4\ W/mm$

(v) $\eta_{ADD} = \frac{P_{OUT}}{P_{DC}}\left(1 - \frac{1}{G}\right) = 27\%$

[30] 30

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Model Answers and Mark Schemes First Examiner: S. LUCYSZYN 11/11

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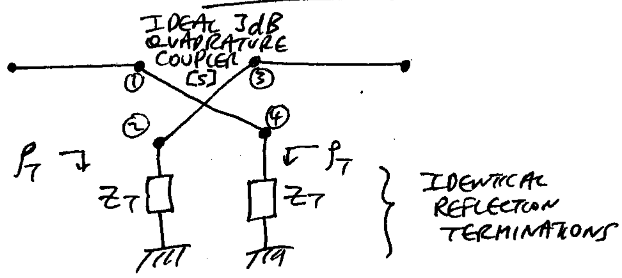
Question Number etc. in left margin | Mark allocation in right margin

SOLUTIONS

6(a)



$$S_{21}|_{OVERALL} = S_{21}\rho_T S_{41} + S_{41}\rho_T S_{21} = 2S_{41}\rho_T S_{21}$$

FOR AN IDEAL 3dB QUADRATURE COUPLER:

$$S_{41} = \frac{1}{\sqrt{2}} e^{-j\pi/2} \quad \text{AND} \quad S_{21} = \frac{1}{\sqrt{2}} e^{j0}$$

$$\therefore S_{21}|_{OVERALL} = 2\frac{1}{\sqrt{2}} e^{-j\pi/2}\rho_T \frac{1}{\sqrt{2}} = \rho_T e^{-j\pi/2}$$

THE REFLECTION-TOPOLOGY TRANSFORMS A REFLECTION COEFFICIENT INTO A TRANSMISSION COEFFICIENT [30] 30

6(b) 6(b) FOR AN ATTENUATOR, THE REFLECTION TERMINATION IS IMPLEMENTED WITH A PIN DIODE OR COLD-FET, TO REALISE A VARIABLE RESISTANCE, R_T.

$$\rho_T(V) = \frac{R_T(V) - Z_0}{R_T(V) + Z_0} \equiv |\rho_T(V)|$$

$$\therefore S_{21}(V)|_{OVERALL} = |\rho_T(V)| e^{-j\pi/2}$$

RELATIVE ATTENUATION, $\Delta|S_{21}(V)|_{OVERALL}| \equiv \Delta|\rho_T(V)|$ [30] 30

6(c) 6(c) FOR A PHASE SHIFTER, THE REFLECTION TERMINATION IS IMPLEMENTED WITH A VARACTOR DIODE, TO REALISE A VARIABLE CAPACITOR, $C_T = -\frac{1}{\omega X_T}$

$$\therefore \rho_T(V) = \frac{jX_T(V) - Z_0}{jX_T(V) + Z_0} \equiv e^{j\angle\rho_T(V)}$$

RELATIVE PHASE SHIFT, $\Delta\angle S_{21}(V)|_{OVERALL} \equiv \Delta\angle\rho_T(V)$

[30] 30

6(d) 6(d) BECAUSE OF UNWANTED PARASITICS, AT HIGH RF FREQUENCIES THE ATTENUATOR SUFFERS FROM AM-PM CONVERSION AND THE PHASE SHIFTER SUFFERS FROM PM-AM CONVERSION. AS A RESULT, EITHER VECTOR CANCELLATION MUST BE APPLIED, AS WITH THE PUSH-PULL SOLUTION, OR A PHASE SHIFTER IN CASCADE WITH AN ATTENUATOR MUST BE USED TO ACHIEVE A USEFUL VECTOR MODULATOR IMPLEMENTATION

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