Final

E4.03
SO10

IMPERIAL COLLEGE LONDON

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2010

MSc and EEE/ISE PART IV: MEng and ACGI

MOBILE RADIO COMMUNICATION

Monday, 17 May 10:00 am

Time allowed: 3:00 hours

There are FOUR questions on this paper.

Answer THREE questions.

All questions carry equal marks. The maximum mark for each subquestion is shown in brackets.

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

Examiners responsible	First Marker(s) :	M.K. Gurcan
	Second Marker(s) :	K.K. Leung

Instructions to Candidates
Useful equations

For knife-edge diffraction model the excess path length

$$\Delta \approx \frac{h^2}{2}\frac{(d_1 + d_2)}{d_1 d_2}$$

and the phase difference

$$\phi \approx \frac{2\pi\Delta}{\lambda} \approx \frac{2\pi}{\lambda}\frac{h^2}{2}\frac{(d_1 + d_2)}{d_1 d_2}$$

Angle

$$\alpha \approx h\frac{(d_1 + d_2)}{d_1 d_2}$$

The diffraction parameter

$$\upsilon = h\sqrt{\frac{2\,(d_1 + d_2)}{\lambda d_1 d_2}} = \alpha\sqrt{\frac{2 d_1 d_2}{\lambda\,(d_1 + d_2)}}$$

Specific Q function values

$$Q\,(-0.4473) = 0.674$$

1. Answer the following sub-questions.

(a) i. Calculate the mean excess delay, rms delay-spread and maximum excess delay for the multipath profile given in figure 1 below. [3]

ii. Estimate the 50% coherence bandwidth of the channel. [3]

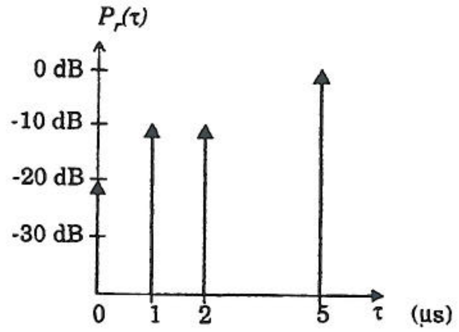

Figure 1. Multipath channel delay profile.

(b) Assume a discrete channel impulse response is used to model urban RF radio channels with excess delays as large as 100 μs and micro-cellular channels with excess delays no longer than 4 μs. If the number of multipath bins is fixed at 64, answer the following sub-questions

i. Find the time resolution $\Delta\tau$ for the impulse response. [2]

ii. Find the maximum RF bandwidth which the two models can accurately represent. [3]

(c) Four received power measurements were taken at distances of 100 m, 200 m, 1 km and 3 km from a transmitter. These measurements are given in the following table.

Distance from transmitter	Received power
100 m	0 dBm
200 m	-20 dBm
1000 m	-35 dBm
3000 m	-70 dBm

Assume that the received power at the reference distance d_0=100m is $P(d_0)$ =0 dBm. Answer the following sub-questions

i. Find the minimum mean square error (mmse) estimate for the path loss exponent, n. [2]

ii. Find the standard deviation, σ, about the mean value. [2]

iii. Estimate the received power at $d = 2$ km using the resulting model. [2]

iv. Predict the likelihood that the received signal level at 2km will be greater than -60 dBm. [2]

v. Predict the percentage of area within a 2 km radius cell that receives signals greater than -60 dBm given in 1.c.iv using figure 2. [2]

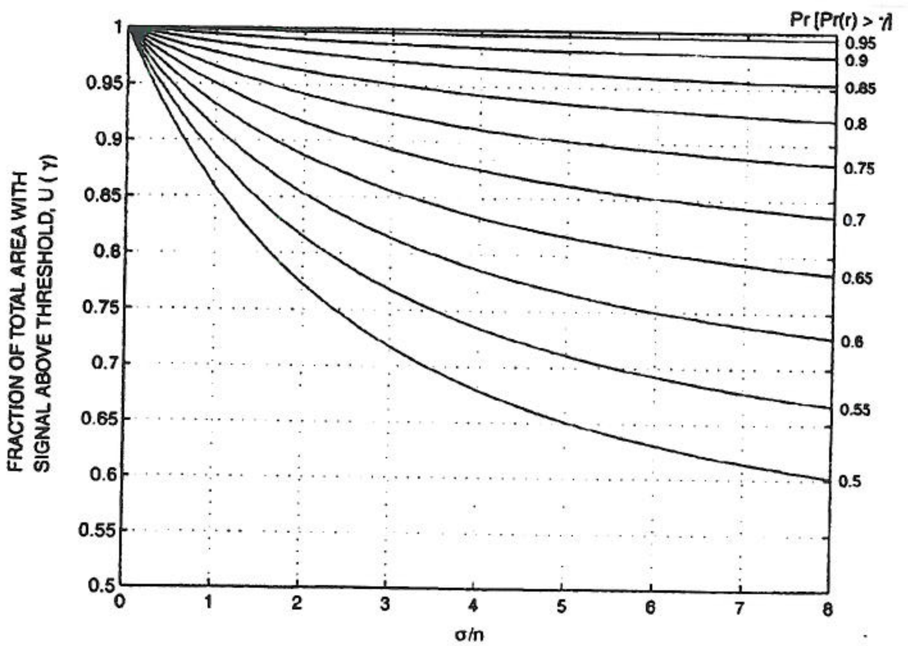

Figure 2. Fraction of total area vs probability of signal above threshold.

(d) Given that an indoor path loss model is of the form

$$PL\,(d)_{dB} = 40 + 20\log d + \sum FAF \quad d \geq 1m$$

where d is measured in metres, find the mean received power between three floors of a building if the floor attenuation factor FAF is 15 dB per floor. Assume that the transmitter radiates 20 dBm and unity gain antennae is used at both transmitter and receiver, and the straight-line path between the transmitter and receiver is 15 m through the floors. [4]

2. Answer the following sub-questions.

(a) Given that a radio system has the transmitter and receiver geometry shown in figure 3 and also that the knife-edge diffraction gain is given as a function of the diffraction parameter v as shown in figure 4,

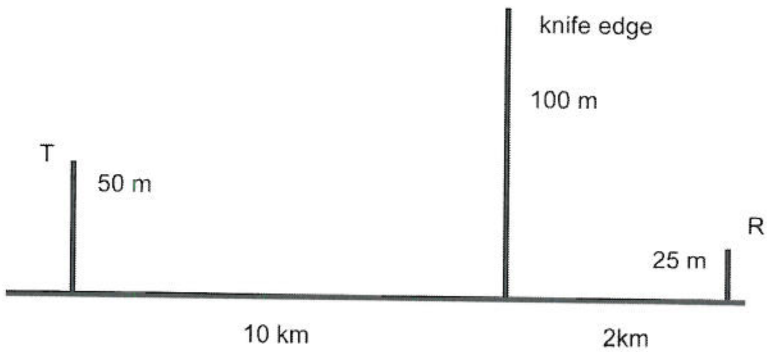

Figure 3. Diffraction geometry

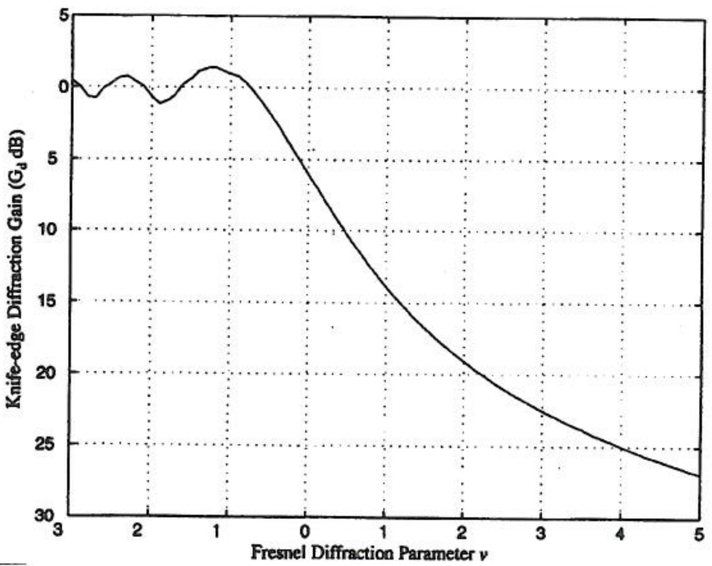

Figure 4. Knife edge diffraction gain

assume that the transmission frequency is f = 900 MHz and answer the following questions.

i. Determine the loss due to knife-edge diffraction. [4]

ii. Determine the height of the obstacle required to introduce 6 dB diffraction loss. [3]

(b) Calculate the diffraction loss for the case shown in figure 5. [3]

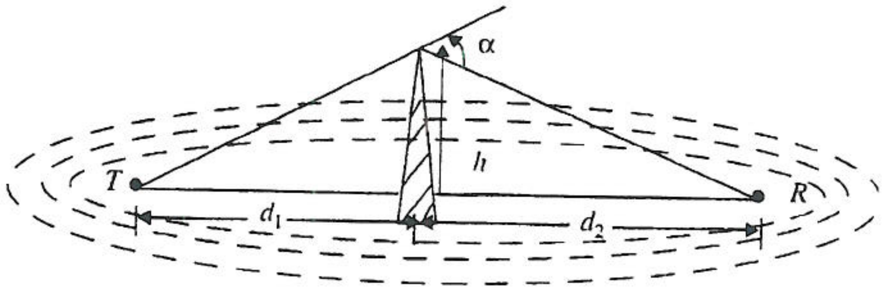

Figure 5. Diffraction geometry

Assume $\lambda = 1/3$ m, $d_1 = 1$ km, $d_2 = 1$ km and $h = 25$ m. Compare your answer using values from figure 4 as well as the appropriate solution given by equations [5]

$$G_d\,(dB) = \begin{cases} 0 & \upsilon \leq -1 \\ 20\log_{10}(0.5 - 0.62\nu) & -1 \leq \upsilon < 0 \\ 20\log_{10}(0.5\exp(-0.95\nu)) & 0 \leq \upsilon < 1 \\ 20\log_{10}\left(0.4 - \sqrt{0.1184 - (0.38 - 0.1\upsilon)^2}\right) & 1 \leq \upsilon < 2.4 \\ 20\log_{10}\left(\dfrac{0.225}{\upsilon}\right) & \upsilon > 2.4 \end{cases}$$

(c) In two-ray path loss model with reflection coefficient $R = -1$, derive an appropriate expression for the location of the signal nulls at the receiver. [6]

(d) Assume that a receiver is located 10 km from a 50 W transmitter. The carrier frequency is 6 GHz and free space propagation is used. The transmitter and receiver gains are $G_t = G_r = 1$ respectively. Find the power at the receiver [4]

3. Answer the following sub-questions.

(a) Consider a flat-fading channel where for a fixed transmit power P, the received SNR is one of four values: $\gamma_1 = 30$ dB, $\gamma_2 = 20$ dB, $\gamma_3 = 10$ dB, and $\gamma_4 = 0$ dB. The probability associated with each state is $p_1 = .2$, $p_2 = .3$, $p_3 = .3$, and $p_4 = .2$. Assume both transmitter and receiver have Channel Side Information.

i. Find the optimal power control policy $P(i)/\bar{P}$ for this channel and its corresponding Shannon capacity per unit Hertz (C/B). [2]

ii. Find the channel inversion power control policy for this channel and associated zero-outage capacity per unit bandwidth. [2]

iii. Find the truncated channel inversion power control policy for this channel and associated outage capacity per unit bandwidth for 3 different outage probabilities: $p_{out} = .1$, $p_{out} = .01$, and p_{out} (and the associated cutoff γ_0) equal to the value that achieves maximum outage capacity. [2]

(b) Assume that the High Speed Downlink Packet Access (HSDPA) system uses K parallel WCDMA channels and the processing gain is $N = 16$ where $K \leq N$. Assume that the expected value of the transmitted symbol is $E\left(v_k^*[x]\, v_k[x]\right) = 1$. The K spread signals are transmitted over the frequency selective channel such that the transmitted signals $\overrightarrow{v}[x]$ are received as being spread by the spreading sequence matrix $\mathbf{HS}$. The received signature sequences for $k = 1, 2, ..., K$ are given by the $(N + 2\alpha)$-length spreading sequence vectors $\vec{q}_k$ for $k = 1, 2, ..., K$ defined by $\mathbf{Q} = [\vec{q}_1, \vec{q}_2, ... \vec{q}_K] = [0^T_{\alpha \times K}, (\mathbf{HS})^T, 0^T_{(\alpha - L+1)\times K}]^T$. The received signal is expressed as:

$$\tilde{r}[x] = \tilde{\mathbf{Q}}\tilde{\mathbf{A}}\tilde{v}[x] + \tilde{n}[x]$$

where $\tilde{n}[x]$ is the noise signal with two-sided noise power spectral density $\frac{N_0}{2}$. The term L is the channel impulse response length. The vector $\widetilde{v}[x]$ is given by $\widetilde{v}[x] = \left[\; \vec{v}_k[x-1]^H,\ \vec{v}_k[x]^H\ \ \vec{v}_k[x+1]^H \;\right]^H$. The receiver signature sequence matrix $\tilde{\mathbf{Q}}$ is defined as $\widetilde{\mathbf{Q}} = \left[\left(\mathbf{J}^T\right)^N \mathbf{Q}, \mathbf{Q}, \mathbf{J}^N\mathbf{Q}\right]$. The matrix $\mathbf{J}$ is the shift matrix. The amplitude matrix $\tilde{\mathbf{A}} = \mathbf{I}_3 \otimes \mathbf{A}$, where $\mathbf{A} = diag(\sqrt{E_1^{(m)}}, \sqrt{E_2^{(m)}}, ..., \sqrt{E_K^{(m)}})$, is used to incorporate the transmission energies $E_k^{(m)}$ for $k = 1, \cdots K$ into the received signal equation. The covariance matrix of the received signals is expressed as

$$\mathbf{C} = E\left(\tilde{r}\,\tilde{r}^H\right) = \tilde{\mathbf{Q}}\tilde{\mathbf{A}}^2\tilde{\mathbf{Q}}^H + N_0\mathbf{I}_{N+2\alpha}.$$

Answer the following sub-questions

i. Show that the receiver despreading filter vector, $\overrightarrow{\omega}_k$, for channel $k = 1, \cdots, K$ is given by [6]

$$\overrightarrow{\omega}_k = \sqrt{E_k}\mathbf{C}^{-1}\vec{q}_k \text{ and}$$

$$\overrightarrow{\omega}_k = \frac{\sqrt{E_k}\left(\mathbf{C} - E_k\,\vec{q}_k\,\vec{q}_k^H\right)^{-1}\vec{q}_k}{1 + E_k\vec{q}_k^H\left(\mathbf{C} - E_k\,\vec{q}_k\,\vec{q}_k^H\right)^{-1}\vec{q}_k}$$

ii. Show that the mean square error, ξ_k^2, for channel $k = 1, \cdots, K$ is given by [3]

$$\xi_k^2 = 1 - \sqrt{E_k}\,\vec{q}_k^H\,\overrightarrow{\omega}_k$$

iii. Show that the signal-to-noise ratio, γ_k, at the output of the k^{th} receiver for channel $k = 1, \cdots, K$ is given by [6]

$$\gamma_k = \frac{E_k\,\vec{q}_k^H\mathbf{C}^{-1}\vec{q}_k}{1 - E_k\,\vec{q}_k^H\mathbf{C}^{-1}\vec{q}_k} \text{ and}$$

$$\gamma_k = E_k\vec{q}_k^H\left(\mathbf{C} - E_k\,\vec{q}_k\,\vec{q}_k^H\right)^{-1}\vec{q}_k$$

iv. Show that the transmitted signal energies $E_k^{(m)}$ for $k = 1, \cdots K$ can be calculated recursively using [4]

$$E_{k,i} = \frac{\gamma_k^*(b_p)}{\gamma_{k,(i-1)}}E_{k,(i-1)}$$

$$= \frac{\Gamma(2^{b_p} - 1)}{\vec{q}_k^H\left(\mathbf{C}_i - E_{k,(i-1)}\,\vec{q}_k\,\vec{q}_k^H\right)^{-1}\vec{q}_k}$$

and

$$E_{k,i}^{(m)} = \frac{\Gamma(2^{b_p} - 1)}{(1 + \Gamma(2^{b_p} - 1))\,\vec{q}_k^H\mathbf{C}_i^{-1}\,\vec{q}_k}$$

where $i = 1, 2, 3, \cdots$ is the iteration number, b_p is the data rate per symbol and Γ is the gap value.

4. Answer the following sub-questions.

(a) Explain the functional differences between the Release 99 and Release 5 Core Networks (CN) for the UMTS system. Explain the functions of the main system entities. [9]

(b) Explain the three step cell search procedure operation for the UTRAN radio system. [8]

(c) When considering the inter-sytem handover from the GSM system to the 3G UTRAN system, using figure 6 explain how the protocols and procedures are used to relocate the Base Station Controllers (BSC) to the 3G networks Radio Network Controller (RNC). [8]

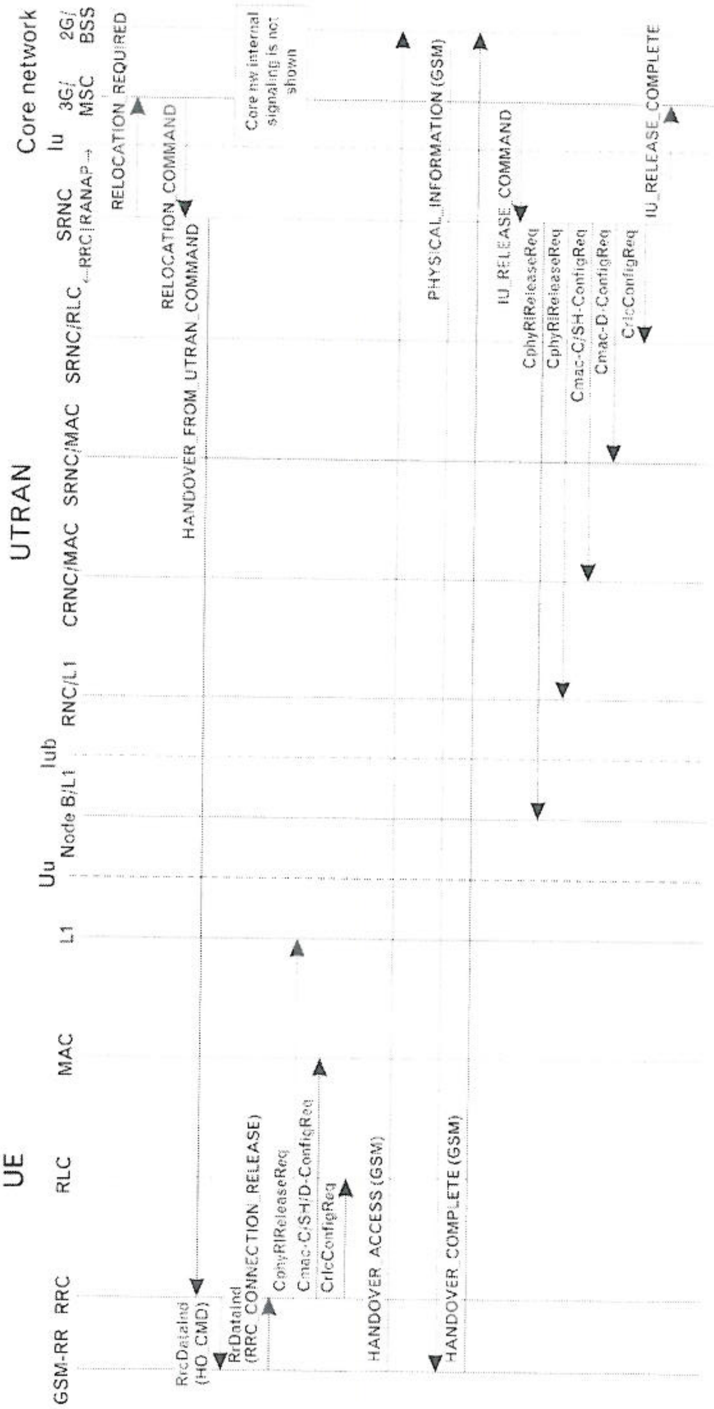

Figure 6. Protocol stack for GSM to UTRAN intersystem handover

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1. (a) The maximum excess delay from 1 is

$\tau_{maximum} = 5\,\mu s.$

The mean excess delay for the given profile is

$$\bar{\tau} = \frac{1 \times 5 + 0.1 \times 1 + 0.1 \times 2 + 0.01 \times 0}{0.01 + 0.1 + 0.1 + 1} = 4.38\,\mu s.$$

$$\overline{\tau^2} = \frac{1 \times 5^2 + 0.1 \times (1)^2 + 0.1 \times (2)^2 + (0.01) \times 0}{1.21} = 21.07\,\mu s^2$$

Therefore rms delay is $\sigma_\tau = \sqrt{21.07 - (4.38)^2}$

$= 1.37\,\mu s$

The coherence bandwidth

$$B_c \approx \frac{1}{5\sigma_\tau} = \frac{1}{5 \times 1.37 \times 10^{-6}} = 146\ kHz.$$

1.b) The maximum excess delay of the channel model is given by $\tau_N = N\Delta\tau$. Therfore $\tau_N = 100\,\mu s$, and $N = 64$ we obtain

$$\Delta\tau = \frac{\tau_N}{N} = 1.5625\,\mu s.$$

The maximum bandwidth is:

$$\frac{2}{\Delta\tau} = \frac{2}{1.5625 \times 10^{-6}} = 1.28\ MHz.$$

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1.(c) (ii)

let P_i be the received power at distance d_i and let $\hat{P}_i$ the estimate of P_i using the $(d/d_0)^n$ path loss model of

$$\overline{PL}(dB) = \overline{PL}(d_0) + 10 \log_{10}\left(\frac{d}{d_0}\right).$$

The sum of squared errors between the measured and estimated values is given by

$$J(n) = \sum_{i=1}^{k} (P_i - \hat{P}_i)^2.$$

The value of n which minimizes the mean square error can be obtained by equating the derivative of $J(n)$ to zero and solving for n. Using $\hat{P}_i = P_i(d_0) - 10 \log_{10}\left(\frac{d_i}{100}\right)^n$. Recognizing that $P_i(d_0) = 0$ dBm we find the following estimates for $\hat{P}_i$ in dBm

$\hat{P}_1 = 0$, $\hat{P}_2 = -3n$, $\hat{P}_3 = -10n$, $\hat{P}_4 = -14.77$

The sum of squared errors is then given by

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1.C.i continued

$J(n) = (0-0)^2 + (-20-(-3n))^2 + (-35-(-10n))^2 + (-70-(-14.77n))^2$

$= 6525 - 2887.8n - \cancel{2887.8}\ 327.153n^2$

$$\frac{dJ(n)}{dn} = 654.306n - 2887.8$$

Setting this equal to zero, the value of n is obtained as $n = 4.4$.

1.c.ii) The sample variance $\sigma^2 = \frac{J(n)}{4}$ at $n = 4.4$ can be obtanied as follows

$J(n) = (0+0) + (-20+13.2)^2 + (-35+44)^2 + (-70+64.988)^2$

$= 152.36$

$\sigma^2 = \frac{152.36}{4} = 38.09 \qquad \sigma = 6.17$

1.c.iii) The probability that the received signal level will be greater than -60 dBm is given by

$$Pr(P_r(d) > -60\,dBm) = Q\left(\frac{\gamma - \overline{P_r(d)}}{\sigma}\right) = Q\left(\frac{-60+57.24}{6.17}\right)$$

$= 67.4\%$

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1.b.ii IF 67.4% OF THE USERS ON THE BOUNDARY RECEIVE SIGNALS GREATER THAN -60 dBm, FROM FIGURE 2 WE CAN DETERMINE

$$\frac{\sigma}{n} = \frac{6.17}{4.4} = 1.402$$

THE CORRESPONDING COVERAGE AREA IS 92%.

1.d Path loss is

$\overline{PL} = 40 + 20 \times \log d + 3 \times 15$

$= 10 + 20 \times \log 15 + 45 = 10 + 45 + 20 \times 1.176$

$= 55 + 23.52 = 78.52$ dB.

Received power

$P_r = 20\,dBm - 78.52\,dB = -58.52\,dBm.$

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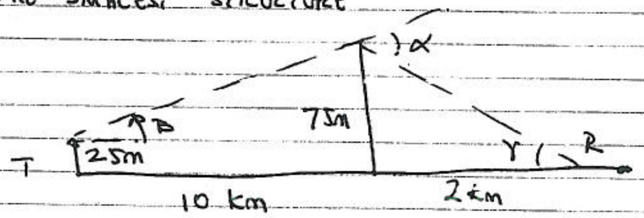
2.a The WAVELENGTH $\lambda = \frac{c}{\lambda} = \frac{3\times10^8}{900\times10^6} = \frac{1}{3}$ m

REDRAW THE GEOMETRY BY SUBTRACTING THE HEIGHT OF THE SMALLEST STRUCTURE



$\beta = \tan^{-1}\left(\frac{75-25}{10000}\right) = 0.2865^\circ$

$\gamma = \tan^{-1}\left(\frac{75}{2000}\right) = 2.15^\circ$ and

$\alpha = \beta + \gamma = 2.434^\circ = 0.0424$ rad

using equation $\nu = \alpha\sqrt{\frac{2d_1 d_2}{\lambda(d_1+d_2)}} = 0.0424\sqrt{\frac{2\times2\times10^6}{\frac{1}{3}(10^4+2\times10^3)}}$

$\nu = 4.24$

FROM $20\log_{10}\left(\frac{0.225}{\nu}\right) = 25.5$ dB.

For 6dB diffraction loss $\nu = 0$

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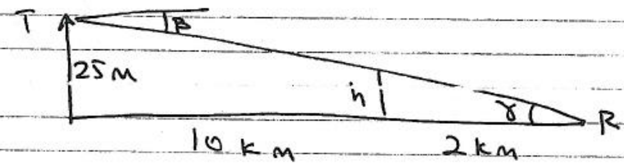
2-b continued

THE OBSTRUCTION HEIGHT h MAY BE FOUND USING TRIANGLES $\beta = \gamma$



It follows THAT

$\frac{h}{2000} = \frac{25}{12000}$, THUS $h = 4.16$ m.

2.c SIGNAL NULLS OCCUR WHEN $\Delta\phi = (2n+1)\pi$

$\frac{2\pi}{\lambda}(x' + x - L) = (2n+1)\pi$

$\frac{2\pi}{\lambda}\left[\sqrt{(h_t+h_r)^2+d^2} - \sqrt{(h_t-h_r)^2+d^2}\right] = \pi(2n+1)$

$\sqrt{(h_t+h_r)^2+d^2} - \sqrt{(h_t-h_r)^2+d^2} = \frac{\lambda}{2}(2n+1)$

let $2n+1 = m$

$\sqrt{(h_t+h_r)^2+d^2} = \frac{m\lambda}{2} + \sqrt{(h_t-h_r)^2+d^2}$

SQUARE BOTH SIDES

$(h_t+h_r)^2+d^2 = m^2\frac{\lambda^2}{4} + (h_t-h_r)^2 + d^2 + m\lambda\left(\sqrt{(h_t-h_r)^2+d^2}\right)$

$X = (h_t+h_r)^2$, $Y = (h_t-h_r)^2$, $X - Y = 4h_t h_r$

$$x = m^2 \frac{\lambda^2}{4} + y + m\lambda\left(\sqrt{y+d^2}\right)$$

$$d = \sqrt{\frac{1}{m\lambda}\left(x - \frac{m^2\lambda^2}{4} - y\right)^2 - y}$$

$$d = \sqrt{\left(\frac{4h_t h_r}{(2n+1)\lambda} - \frac{(2n+1)\lambda}{4}\right)^2 - (h_t - h_r)^2}$$

2d

$d = 10^4$ m

$P_T = 50$ w

$f_c = 6.\times10^9$ Hz

$$\lambda = \frac{c}{f} = \frac{3\times10^8}{6\times10^9} = 0.05$$

$$P_r = \left(\frac{\sqrt{G}\,\lambda}{4\pi d}\right)^2 P_t = \left(\frac{5\times10^{-2}}{4\pi\times10^4}\right) 50 \text{ w.}$$

3. a) WE SUPPOSE THAT ALL CHANNEL STATES ARE USED.

i.) $\frac{1}{\gamma} = 1 + \sum_{i=1}^{4} \frac{1}{\gamma_i} P_i \Rightarrow \gamma_0 = 0.8109$

$\frac{1}{\gamma_0} - \frac{1}{\gamma_4} > 0$ ∴ true as γ_4 is the smallest.

$$\frac{P(\gamma_i)}{\bar{P}} = \frac{1}{\gamma_0} - \gamma_i$$

$$\frac{P(\gamma)}{\bar{P}} = \begin{cases} 1.2322 & \gamma = \gamma_1 \\ 1.2232 & \gamma = \gamma_2 \\ 1.1322 & \gamma = \gamma_3 \\ 0.2332 & \gamma = \gamma_4 \end{cases}$$

$$\frac{C}{B} = \sum_{i=1}^{4} \log_2\left(\frac{\gamma_i}{\gamma_0}\right) p(\gamma_i) = 5.2853 \text{ bps/Hz}$$

ii) $\sigma = \frac{1}{E(\frac{1}{\gamma})} = 4.2882$

$$\frac{P(\gamma_i)}{\bar{P}} = \frac{\sigma}{\gamma_i}, \quad \frac{P(\gamma)}{\bar{P}} = \begin{cases} 0.0043 & \gamma = \gamma_1 \\ 0.0029 & \gamma = \gamma_2 \\ 0.4288 & \gamma = \gamma_3 \\ 4.2882 & \gamma = \gamma_4 \end{cases}$$

$$\frac{C}{B} = \log_2(1+\sigma) = 2.4028 \text{ bps/Hz}$$

3.a. iii)

TO HAVE $P_{OUT} = 0.1$ or 0.01 WE WILL HAVE TO USE ALL THE SUB-CHANNELS AS LEAVING ANY OF THESE WILL RESULT IN P_{OUT} OF AT LEAST 0.2 ∴ TRUNCATED CHANNEL POWER CONTROL POLICY AND ASSOCIATED SPECTRAL EFFICIENCY ARE THE SAME AS ZERO-OUTAGE CASE IN 3.a.ii. TO HAVE P_{OUT} THAT MAXIMIZES C WITH TRUNCATED CHANNEL INVERSION WE GET

$$\text{Max.} \frac{C}{B} = 4.1462 \ \text{bps/Hz} \qquad P_{OUT} = 0.5$$

—///—

3.b. i)

$$E\left| v_k - w_k^H \tilde{r} \right|^2, \quad \text{where} \quad E(v_k^* \tilde{r}) = \sqrt{E_k}\, \vec{q}_k$$

$$E(v_k^* v_k) = 1$$

The received vector satisfies the relationship

$$E(v_k r - \bar{r}\bar{r}^H \vec{w}_k) = 0$$

Hence

$$\vec{w}_k = E[r r^H]^{-1} E(v_k^* \tilde{r}) = \sqrt{E_k}\, C^{-1} \vec{q}_k$$

—//—

$$\therefore \mathcal{E}_k^2 = 1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k$$

—///—

$$\gamma_k = \frac{1}{\mathcal{E}_k^2} - 1 = \frac{\sqrt{E_k}\, \vec{q}_k^H \vec{w}_k}{1 - \sqrt{E}\, \vec{q}_k^H \vec{w}_k}$$

$$\gamma_k = \frac{E_k \left| \vec{q}_k^H \vec{w}_k \right|^2}{\vec{w}_k^H (C - E_k \vec{q}_k \vec{q}_k^H) \vec{w}_k}$$

$$\frac{\sqrt{E_k}\, \vec{q}_k^H \vec{w}_k}{1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k} = \frac{E_k \left| \vec{q}_k^H \vec{w}_k \right|^2}{\vec{w}_k^H (C - E_k \vec{q}_k \vec{q}_k^H) \vec{w}_k}$$

$$\sqrt{E_k}\, \vec{q}_k^H \vec{w}_k = \sqrt{E_k}\, \cancel{\vec{w}_k^H} \vec{q}_k = \frac{\cancel{\vec{w}_k^H} (C - E_k \vec{q}_k \vec{q}_k^H) \vec{w}_k}{1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k}$$

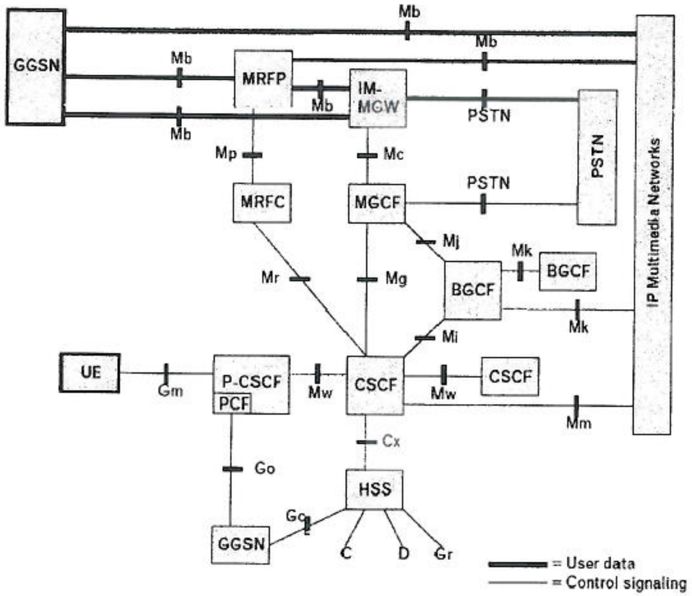
$$\vec{w}_k = \sqrt{E_k} \left(1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k\right) \left(C - E_k \vec{q}_k \vec{q}_k^H\right)^{-1} \vec{q}_k$$

$$\gamma_k = \frac{\sqrt{E_k}\, \vec{q}_k^H \left(C - E_k \vec{q}_k \vec{q}_k^H\right)^{-1} \vec{q}_k \, \cancel{(1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k)}}{\cancel{(1 - \sqrt{E_k}\, \vec{q}_k^H \vec{w}_k)}}$$

$$\boxed{\gamma_k = \sqrt{E_k}\sqrt{E_k}\, \vec{q}_k^H \left(C - E_k \vec{q}_k \vec{q}_k^H\right)^{-1} \vec{q}_k}$$

$$\vec{w}_k = \sqrt{E_k}\, \frac{1}{1 + \gamma_k} \left(C - E_k \vec{q}_k \vec{q}_k^H\right)^{-1} \vec{q}_k$$

$$= \sqrt{E_k} \left(C - E_k \vec{q}_k \vec{q}_k^H\right) \vec{q}_k$$



To make the high-speed data transfer more efficient, IMS has a new approach to the network design. Previously, data was transferred through several network elements on its way to its destination. In the new system, data typically bypasses the control logic in the core network. The old CS switch, MSC, has been divided into two logical entities, a media gateway (MGW), and an MSC server (MSC). The control logic is in MSC, and the actual switching matrix in MGW. These logical entities can be implemented in the same or separate physical units. The separation of control and data traffic enables the network to employ more efficient routers for the high-speed data, as the small-sized control messages are handled elsewhere.

An *All-IP network* means that all traffic data, including voice, is transferred as IP packets. This opens the 3G mobile environment to the large IP applications industry. The problem of mobile networks has been to find revenue-generating applications, and IP Multimedia Domain will make this task easier. The new applications do not necessarily have to be developed for the mobile environment anymore, at least not because of the transport technique used. Another important argument for All-IP networks is that the technology makes the separation of PS and CS domains obsolete. All-IP networks make the transport technology uniform, and that should reduce network-deployment costs. Voice can also be handled as packets in an All-IP network. Note that VoIP as such is hardly an improvement for voice transfer. Circuit-switched systems were originally designed for voice transfer; they can do it quite efficiently, and provide high-quality results. However, All-IP networks bring lots of advantages; thus, voice too has to be transformed into a packet service.

The problem from the network point of view is that it has to be backwards-compatible with earlier releases. There will be lots of Release 99 devices in use when Release 5 is deployed, and those would become useless if there were no backwards compatibility. A UE can use IMS domain services if it has

Primary scrambling code

$$C_{PSC} = (1+j) \times (a, a, a, -a, -a, a, -a, -a, a, a, a, -a, a, -a, a, a)$$

$$a = (x_1, x_2, x_3, \cdots, x_{16})$$
$$= (1, 1, 1, 1, 1, 1, -1, -1, 1, -1, 1, -1, 1, -1, -1, 1)$$

Secondary scrambling codes

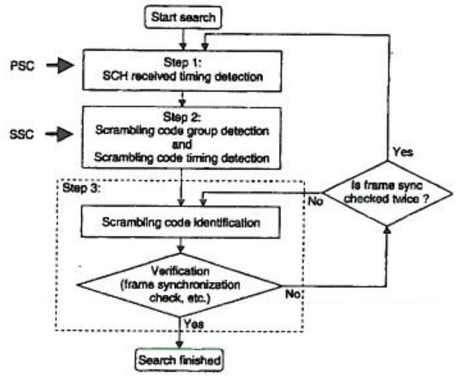
$$n = 16 \times (k-1)$$

H_8 Hadamar codes

$$C_{SSC,k} = (1+j) \times (h_n(0) \times z(0), h_n(1) \times z(1), h_n(2) \times z(2), \cdots \times h_n(255) \times z(255))$$

$$z = (b, b, b, -b, b, b, -b, -b, b, -b, b, -b, -b, -b, -b, -b)$$

$$b = (1, 1, 1, 1, 1, 1, 1, 1, -1, -1, -1, -1, -1, -1, -1, -1)$$



Group code identification

Scrambling Code Group	slot number														
	#0	#1	#2	#3	#4	#5	#6	#7	#8	#9	#10	#11	#12	#13	#14
Group 0	1	1	2	8	9	10	15	8	10	16	2	7	15	7	16
Group 1	1	1	5	16	7	3	14	16	3	10	5	12	14	12	10
Group 2	1	2	1	15	5	5	12	16	6	11	2	16	11	15	12
Group 3	1	2	3	1	8	6	5	2	5	8	4	4	6	3	7
Group 4	1	2	16	6	6	11	15	5	12	1	15	12	16	11	2
Group 5	1	3	4	7	4	1	5	5	3	6	2	8	7	6	8

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