1.

IIA 3F4 Data Transmission

1.) (a) For binary PAM with transmitted levels A_1 and A_2 we have

$$V_1 - V_0 = (A_2 - A_1)\, h_0 = (A_2 - A_1)\, k\, p_R(0)$$

We must maximise $\left(\frac{V_1 - V_0}{2\sigma_v}\right) = \frac{(A_2 - A_1)\, k\, p_R(0)}{2\sigma_v}$.

For a given system A_1, A_2 and $p_R(0)$ are fixed hence we must maximise k/σ_v or equivalently minimise $\frac{\sigma_v}{k}$ (or $\frac{\sigma_v^2}{k^2}$) 1

The psd of the received noise at the slicer is

$$S_v(\omega) = N(\omega)\, |H_R(\omega)|^2$$

hence the noise power at the slicer is,

$$\sigma_v^2 = \frac{1}{2\pi}\int_{-\infty}^{\infty} S_v(\omega)\, d\omega = \frac{1}{2\pi}\int_{-\infty}^{\infty} N(\omega)\, |H_R(\omega)|^2\, d\omega.$$ 1

Now express gain term k in terms of the transmitted pulse $h_T(t)$.

$$E_T = \int_{-\infty}^{\infty} (h_T(t))^2\, dt$$

From Parsevals theorem,

$$E_T = \frac{1}{2\pi}\int_{-\infty}^{\infty} |H_T(\omega)|^2\, d\omega.$$ 1

We know,

$$H_T(\omega)\, H_C(\omega)\, H_R(\omega) = k\, P_R(\omega), \text{ so}$$

$$H_T(\omega) = \frac{k\, P_R(\omega)}{H_C(\omega)\, H_R(\omega)} \quad \text{giving}$$

$$E_T = \frac{1}{2\pi}\int_{-\infty}^{\infty} \frac{k^2\, |P_R(\omega)|^2}{|H_C(\omega)|^2\, |H_R(\omega)|^2}\, d\omega, \text{ so}$$

$$k^2 = \frac{E_T}{\frac{1}{2\pi}\int \frac{|P_R(\omega)|^2}{|H_C(\omega)|^2\, |H_R(\omega)|^2}\, d\omega}$$ 1

2.

So,

$$\frac{\sigma_v^2}{k^2} = \frac{1}{(2\pi)^2 E_T}\int_{-\infty}^{\infty} N(\omega)\,|H_R(\omega)|^2 d\omega \int_{-\infty}^{\infty} \frac{|P_R(\omega)|^2}{|H_c(\omega)|^2\,|H_R(\omega)|^2}\, d\omega \qquad 1$$

Schwartz inequality states that

$$\int_{-\infty}^{\infty} |F(\omega)|^2 d\omega \int |G(\omega)|^2 d\omega \geq \left| \int_{-\infty}^{\infty} F(\omega)\,G(\omega)\, d\omega \right|^2$$

with equality when $F(\omega) = \lambda\, G^*(\omega)$ where λ is an arbitrary constant.

Let,

$$F(\omega) = \sqrt{N(\omega)}\;|H_R(\omega)| \quad \text{and}$$

$$G(\omega) = \frac{|P_R(\omega)|}{|H_c(\omega)|\,|H_R(\omega)|}$$

So we obtain,

$$\frac{\sigma_v^2}{k^2} \geq \frac{1}{(2\pi)^2 E_T} \left| \int_{-\infty}^{\infty} \sqrt{N(\omega)}\; \frac{|P_R(\omega)|}{|H_c(\omega)|}\, d\omega \right|^2 \qquad 1$$

All terms on rhs are fixed, hence ~~$\frac{\sigma_v^2}{k^2}$~~ $\frac{\sigma_v^2}{k^2}$ is minimised when two sides of eqn are equal, ie $F(\omega) = \lambda\, G^*(\omega)$

ie,

$$\sqrt{N(\omega)}\;|H_R(\omega)| = \lambda\, \frac{|P_R(\omega)|}{|H_c(\omega)|\,|H_R(\omega)|}$$

Since λ is arbitrary, let $\lambda = 1$, so

$$|H_R(\omega)| = \left| \frac{P_R(\omega)}{\sqrt{N(\omega)}\;H_c(\omega)} \right|^{1/2} \qquad 1$$

and for ~~wht~~ $N(\omega) = N_0$,

$$|H_R(\omega)| = \left| \frac{P_R(\omega)}{\sqrt{N_0}\;H_c(\omega)} \right|^{1/2} \quad \text{ie} \quad \propto \left| \frac{P_R(\omega)}{H_c(\omega)} \right|^{\frac{1}{2}}$$

similarly,

$$|H_T(\omega)| = k \left| \frac{P_R(\omega)\sqrt{N(\omega)}}{H_c(\omega)} \right|^{1/2}$$

for, $N(\omega) = N_0$,

$$|H_T(\omega)| = k \left| \frac{P_R(\omega)\sqrt{N_0}}{H_c(\omega)} \right|^{1/2} \qquad 1$$

$$\text{ie} \quad \propto \left| \frac{P_R(\omega)}{H_c(\omega)} \right|^{1/2}$$

[8]

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(b)(i) $$|H_R(\omega)| = \left| \frac{P_R(\omega)}{\sqrt{N_0}\ H_c(\omega)} \right|^{1/2}$$

and

$$|H_T(\omega)| = k \left| \frac{P_R(\omega)\sqrt{N_0}}{H_c(\omega)} \right|^{1/2}$$

Substitute in for $H_c(\omega)$ and $P_R(\omega)$ gives,

$$|H_R(\omega)| = \left| \frac{T\cos^2\left(\frac{\omega T}{2}\right)}{\sqrt{N_0}\ \cos\left(\frac{\omega T}{2}\right)} \right|^{1/2}$$

$$= \left| \frac{T}{\sqrt{N_0}} \cdot \cos\left(\frac{\omega T}{2}\right) \right|^{1/2} \quad \text{for } |\omega| \le \pi/T$$

$$= \sqrt{\frac{T}{\sqrt{N_0}} \cdot \cos\left(\frac{\omega T}{2}\right)} \qquad 2$$

and,

$$|H_T(\omega)| = k \left| \frac{\sqrt{N_0}\, T\cos^2\left(\frac{\omega T}{2}\right)}{\cos\left(\frac{\omega T}{2}\right)} \right|^{1/2}$$

$$= k\sqrt{\sqrt{N_0}\, T\cos\left(\frac{\omega T}{2}\right)} \quad \text{for } |\omega| \le \pi/T.$$

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[4]

(ii) Now the PSD of the unity magnitude random impulse train is $\frac{1}{T}$.

Consequently the signal power at the slicer input is,

$$P_{Rx} = \frac{1}{2\pi}\int_{-\infty}^{\infty} \frac{1}{T} \cdot |H_T(\omega)|^2 |H_c(\omega)|^2 |H_R(\omega)|^2 \, d\omega \qquad 1$$

$$P_{Rx} = \frac{1}{2\pi T}\int_{-\pi/T}^{\pi/T} \left(k\sqrt{\sqrt{N_0}\, T\cos\left(\frac{\omega T}{2}\right)}\right)^2 \cdot \cos^2\left(\frac{\omega T}{2}\right) \cdot \left(\sqrt{\frac{T}{\sqrt{N_0}} \cdot \cos\left(\frac{\omega T}{2}\right)}\right)^2 d\omega$$

$$P_{Rx} = \frac{1}{2\pi T}\int_{-\pi/T}^{\pi/T} k^2\sqrt{N_0}\, T\cos\left(\frac{\omega T}{2}\right) \cos^2\left(\frac{\omega T}{2}\right) \cdot \frac{T}{\sqrt{N_0}} \cos\left(\frac{\omega T}{2}\right) d\omega$$

4

$$P_{Rx} = \frac{1}{2\pi T}\int_{-\frac{\pi}{T}}^{\pi/T} k^2 T^2 \cos^4\left(\frac{wT}{2}\right)\, dw$$

$$P_{Rx} = \frac{k^2 T}{2\pi}\int_{-\pi/T}^{\pi/T} \cos^4\left(\frac{wT}{2}\right)\, dw$$

$$P_{Rx} = \frac{k^2 T}{2\pi}\int_{-\pi/T}^{\pi/T} \cos^2\left(\frac{wT}{2}\right)\cos^2\left(\frac{wT}{2}\right)\, dw$$

$$= \frac{k^2 T}{2\pi}\int_{-\pi/T}^{\pi/T} \frac{(1+\cos wT)}{2}\,\frac{(1+\cos wT)}{2}\, dw$$

$$= \frac{k^2 T}{8\pi}\int_{-\pi/T}^{\pi/T} 1 + 2\cos wT + \cos^2 wT\, dw$$

$$= \frac{k^2 T}{8\pi}\int_{-\pi/T}^{\pi/T} 1 + 2\cos wT + \frac{(1+\cos 2wT)}{2}\, dw$$

$$= \frac{k^2 T}{8\pi}\int_{-\pi/T}^{\pi/T} \frac{3}{2} + 2\cos wT + \frac{1}{2}\cos 2wT\, dw$$

$$= \frac{k^2 T}{8\pi}\left[\frac{3}{2}w + \frac{2}{T}\sin wT + \frac{1}{2\cdot 2T}\sin 2wT\right]_{-\pi/T}^{\pi/T}$$

$$= \frac{k^2 T}{8\pi}\left[\left[\frac{3}{2}\cdot\frac{\pi}{T} + \frac{2}{T}\sin\frac{\pi}{T}\cdot T + \frac{1}{4T}\sin\frac{2T\pi}{T}\right] - \left[\frac{-3}{2}\frac{\pi}{T} + \frac{2}{T}\sin\frac{-\pi}{T}\cdot T + \frac{1}{4T}\sin -\frac{2T\pi}{T}\right]\right]$$

$$= \frac{k^2 T}{8\pi}\left[\frac{3\pi}{2T} + \frac{3\pi}{2T}\right] = \frac{k^2 T}{8\pi}\left(\frac{6\pi}{2T}\right) = \underline{\underline{\frac{3k^2}{8}}} \qquad 1$$

Now work out the noise power at the slicer input.

$$P_{NR} = \frac{1}{2\pi}\int_{-\pi/T}^{\pi/T} N_o \; |H_R(\omega)|^2 \, d\omega$$ 1

$$P_{NR} = \frac{N_o}{2\pi}\int_{-\pi/T}^{\pi/T} \frac{T}{\sqrt{N_o}} \cos\left(\frac{\omega T}{2}\right) d\omega$$

$$= \frac{T\sqrt{N_o}}{2\pi}\int_{-\pi/T}^{\pi/T} \cos\left(\frac{\omega T}{2}\right) d\omega$$

$$= \frac{\cancel{T}\sqrt{N_o}}{\cancel{2}\pi} \cdot \frac{\cancel{2}}{\cancel{T}}\left[\sin\left(\frac{\omega T}{2}\right)\right]_{-\pi/T}^{\pi/T}$$

$$= \frac{\sqrt{N_o}}{\pi}\left[\left(\sin \frac{\pi}{T}\cdot\frac{T}{2}\right) - \sin\left(-\frac{\pi}{T}\cdot\frac{T}{2}\right)\right]$$

$$= \frac{\sqrt{N_o}}{\pi}\left[\sin\frac{\pi}{2} - \sin\left(\frac{-\pi}{2}\right)\right]$$

$$= \frac{2\sqrt{N_o}}{\pi}$$ 1

$$SNR = \frac{P_{Rx}}{P_{NR}} = \frac{3k^2}{8}\cdot\frac{\pi}{2\sqrt{N_o}}$$

$$= \frac{3\pi k^2}{16\sqrt{N_o}}$$ 1

Power at output of transmit filter is,

$$P_{Tx} = \frac{1}{2\pi}\int_{-\pi/T}^{\pi/T} \frac{1}{T} \; |H_T(\omega)|^2 \, d\omega$$ 1

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$$P_{Tx} = \frac{1}{2\pi T} \int_{-\pi/T}^{\pi/T} k^2 \sqrt{N_0}\, T \cos^2\left(\frac{\omega T}{2}\right) d\omega$$

$$P_{Tx} = \frac{k^2 \sqrt{N_0}\, \cancel{T}}{2\pi \cancel{T}} \int_{-\pi/T}^{\pi/T} \cos^2\left(\frac{\omega T}{2}\right) d\omega$$

$$P_{Tx} = \frac{k^2 \sqrt{N_0}}{2\pi} \cdot 2 \cdot \frac{2}{T} = \frac{2 k^2 \sqrt{N_0}}{\pi T} \qquad 1$$

So, $$\frac{SNR}{P_{Tx}} = \frac{3\pi k^2}{16\sqrt{N_0}} \cdot \frac{\pi T}{2 k^2 \sqrt{N_0}}$$

$$= \underline{\underline{\frac{3\pi^2 T}{32 N_0}}} \qquad 1$$

[8]

2.) (a) Display the signal at the input to the data slicer on an oscilloscope triggered using the symbol clock - set the time base to show 1 or 2 symbol periods.

The 'openness' of the eye gives guidance concerning the ease of making correct decisions. The height of the eye opening indicates a lower probability of bit error.
The width indicates susceptibility to symbol timing errors.

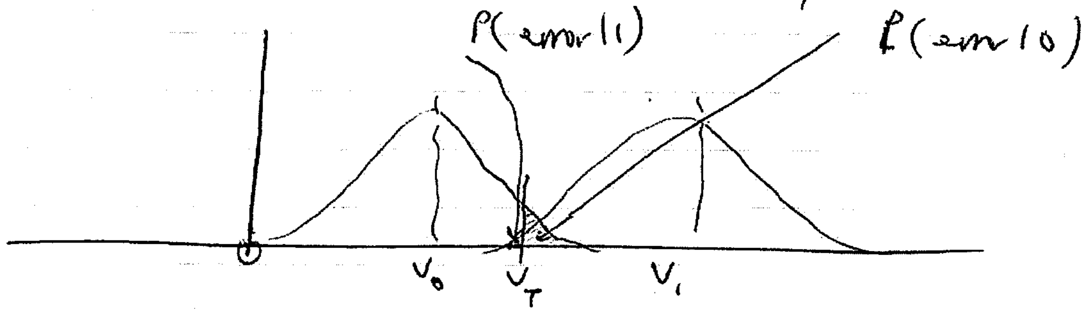
When used to estimate BER then it is normal ~~you have~~ to assume worst case opening - however will not always be so closed.

[4]

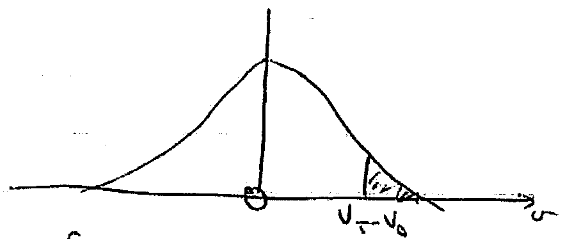
(b) Total BER is

$P_e = P(\text{error}|0)\, P_0 + P(\text{error}|1)\, P_1$ 1

now need to work out cond error prob.



Consider $P(\text{error}|0)$, redraw as follows



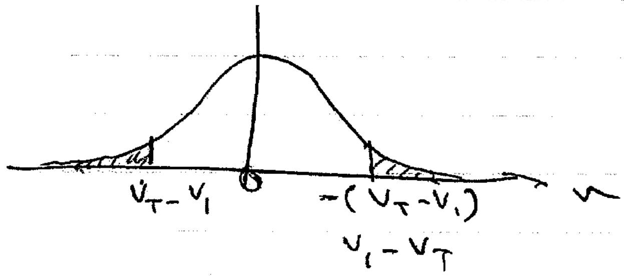
For a Gaussian dist with zero mean and variance σ^2

$$P(\text{error}|0) = \frac{1}{\sqrt{2\pi\sigma^2}} \int_{V_T - V_0}^{\infty} e^{-\frac{v^2}{2\sigma^2}}\, dv$$

Rewriting as Q function

$$P(\text{error}|0) = \frac{1}{\sqrt{2\pi}} \int_{\frac{V_T - V_0}{\sigma}}^{\infty} e^{-\frac{v^2}{2}}\, dv = Q\left(\frac{V_T - V_0}{\sigma}\right) \quad 1$$

Consider $P(\text{error}|1)$



$$P(\text{error}|1) = \frac{1}{\sqrt{2\pi\sigma^2}} \int_{V_1 - V_T}^{\infty} e^{-\frac{v^2}{2\sigma^2}}\, dv$$

$$P(\text{error}|1) = \frac{1}{\sqrt{2\pi}} \int_{\frac{V_1 - V_T}{\sigma}}^{\infty} e^{-\frac{v^2}{2}}\, dv \quad 1$$

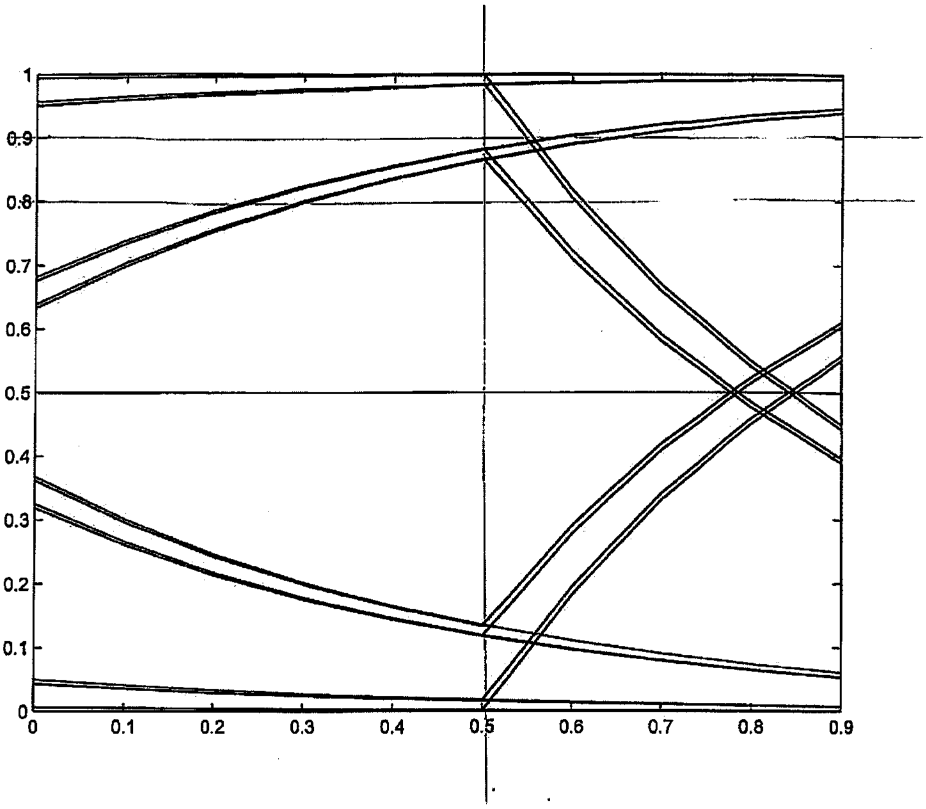
$$P(\text{error}|1) = Q\left(\frac{V_1 - V_T}{\sigma}\right) \quad 1$$

So,

$$P_e = P_0\, Q\left(\frac{V_T - V_0}{\sigma}\right) + P_1\, Q\left(\frac{V_1 - V_T}{\sigma}\right). \quad 1$$

[6]

(c)(i)



[4]

sketch should show at least max and min eye openings.

(ii) Assumptions

- Worst case (min eye opening)
- Optimum sampling instant
- $P_0 = P_1 = 0.5$. 1

The worst case '1' occurs when a single '1' follows a run of '0's. If the input is considered as a step of amplitude A, then the output of the RC channel is

$$V_0 = A\left(1 - e^{-t/RC}\right)$$

when the bit period $T = 2RC$, then at ideal sample instant,

$$V_0 = A\left(1 - e^{-2RC/RC}\right)$$
$$= A\left(1 - e^{-2}\right)$$
$$= 0.865\,A.$$

For $A = 1V$,

$V_0 = 0.865V$ 1

Worst case '0' is when a single '0' follows a run of '1's. In this case the output of the channel is given by

$V_0 = A e^{-t/RC}$

assuming C is initially charged to A V.
In the example, $A = 1V$ and the bit period $t = 2RC$, so

$V_0 = e^{-2RC/RC}$

$= e^{-2}$

$= 0.135V$ 1

So,

$P(\text{error}|0) = Q\left(\frac{V_T - V_0}{\sigma_v}\right)$

$= Q\left(\frac{0.4 - 0.135}{0.1}\right) = Q(2.65)$

$= 4.03 \times 10^{-3}$

$P(\text{error}|1) = Q\left(\frac{V_1 - V_T}{\sigma_v}\right)$

$= Q\left(\frac{0.865 - 0.4}{0.1}\right) = Q(4.65)$

$= 1.66 \times 10^{-6}$

So,

$$BER = (0.5 \times 4.03 \times 10^{-3}) + (0.5 \times 1.66 \times 10^{-5})$$
$$= \underline{\underline{2.01 \times 10^{-3}}} \qquad 1$$

i.e., dominated by errors detecting '0's. [4]

(ii) Optimum threshold voltage, $V_{T,opt}$

$$V_{T,opt} = \frac{V_0 + V_1}{2} = \frac{0.865 - 0.135}{2} = \underline{\underline{0.5}} \qquad 1$$

Also assume $P_0 = P_1 = 1/2$,
sub for $V_T = V_{T,opt}$ and $P_0 = P_1 = 1/2$ in answer to (b),

$$P_e = \frac{1}{2} Q\left(\left(\frac{V_0 + V_1}{2} - V_0\right) \cdot \frac{1}{\sigma}\right) + \frac{1}{2} Q\left(\left(V_1 - \left(\frac{V_0 + V_1}{2}\right)\right)\frac{1}{\sigma}\right)$$

$$= \frac{1}{2} Q\left(\frac{1}{\sigma}\left(\frac{V_1 - V_0}{2}\right)\right) + \frac{1}{2} Q\left(\frac{1}{\sigma}\left(\frac{V_1 - V_0}{2}\right)\right)$$

$$= Q\left(\frac{V_1 - V_0}{2\sigma}\right) = \frac{h}{2\sigma}$$

where h is worst case eye opening.

$$h = 0.865 - 0.135 = 0.729$$

$$\therefore P_e = Q\left(\frac{h}{2\sigma}\right)$$

$$= Q\left(\frac{0.729}{2 \times 0.1}\right) = Q(3.65) \qquad 1$$

$$= \underline{\underline{1.31 \times 10^{-4}}}$$

[2]

3.(a)

$$s(t) = \mathrm{Re}\left(p(t)e^{j\omega_c t}\right)$$
$$= \mathrm{Re}\left(\left(u(t) + j\,v(t)\right)\left(\cos\omega_c t + j\sin\omega_c t\right)\right)$$
$$= u(t)\cos\omega_c t - v(t)\sin\omega_c t$$

To recover $u(t)$, we multiply $s(t)$ by $2\cos\omega_c t$:

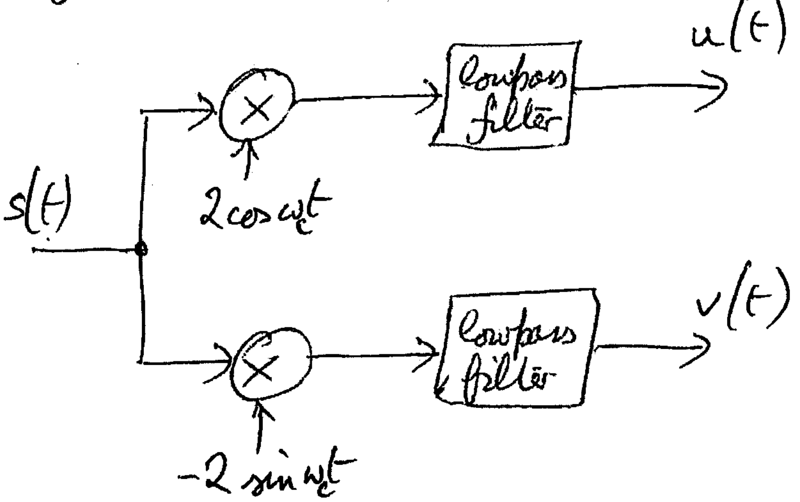
$$\therefore\; s(t).2\cos\omega t = u(t).\,2\cos^2\omega_c t - v(t).\,2\sin\omega_c t\cos\omega_c t$$
$$= u(t)\left(1+\cos 2\omega_c t\right) - v(t).\,\sin 2\omega_c t$$

Applying a lowpass filter to remove the $\cos 2\omega_c t$ & $\sin 2\omega_c t$ modulated terms gives just $u(t)$.

Similarly to recover $v(t)$, multiply $s(t)$ by $(-2\sin\omega_c t)$:

$$\therefore\; s(t).(-2\sin\omega_c t) = u(t)\left(-2\cos\omega_c t\sin\omega_c t\right) + v(t).\,2\sin^2\omega_c t$$
$$= u(t).(-\sin 2\omega_c t) + v(t)\left(1-\cos 2\omega_c t\right)$$

Apply a lowpass filter here gives $v(t)$.



[4]

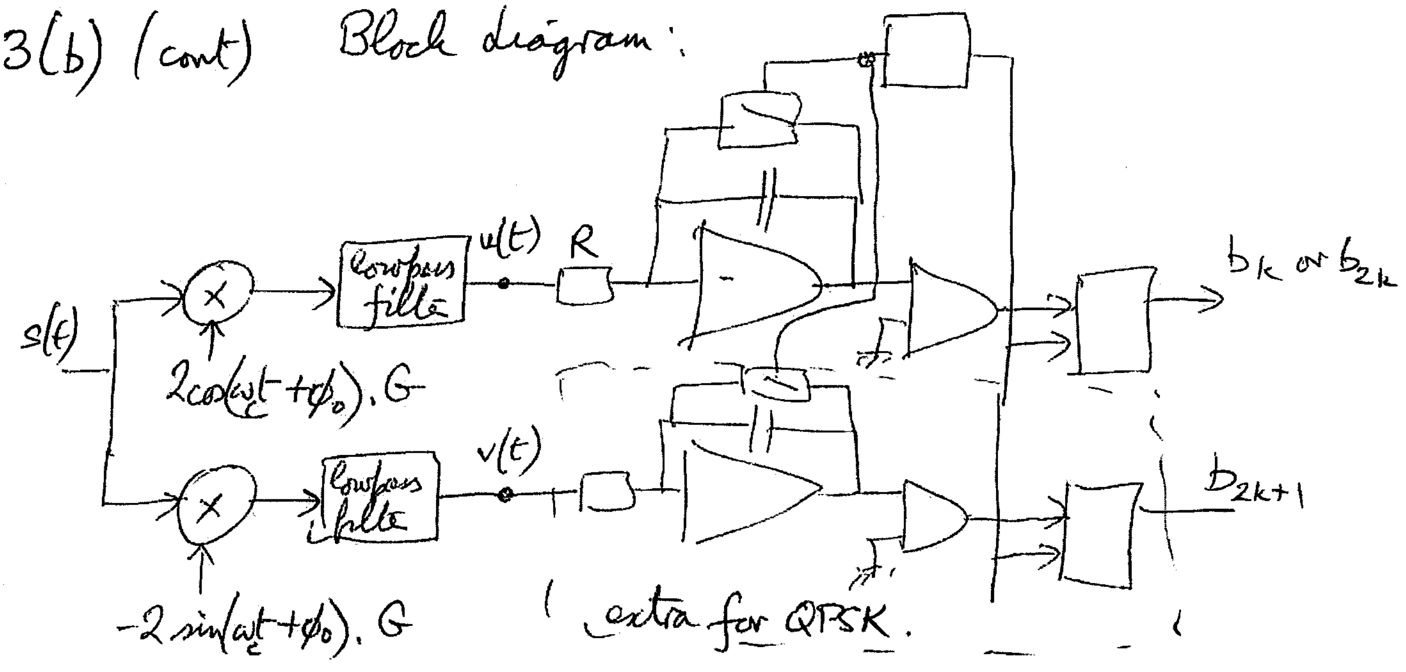
3(b) This system calculates

$$y(k) = \int_{kT_b}^{(k+1)T_b} Re\left[r(t).\, G g(t-kT_b)\, e^{-j\phi_0}\right] dt$$

at the output of the matched filter. It samples the polarity of $y(k)$ at the end of the k^{th} bit period at time $(k+1)T_b$ using the Threshold Detector and the Sampling Register to generate the demodulated output bit b_k. $r(t)$ is the received complex phasor waveform, $g(t)$ is the shaping pulse used for each bit at the transmitter, G is an arbitrary gain constant, and ϕ_0 is the phase shift in the demodulator which compensates for the phase offset of the received signal $r(t)$ due to delay on the signal path.

The monostable and electronic switch are designed to discharge the capacitor C at the start of each bit period (i.e. at kT_b for all integers k), so that the above integration starts afresh at kT_b, for each k. The pulse-width and discharge time must be short compared with the bit period T_b.

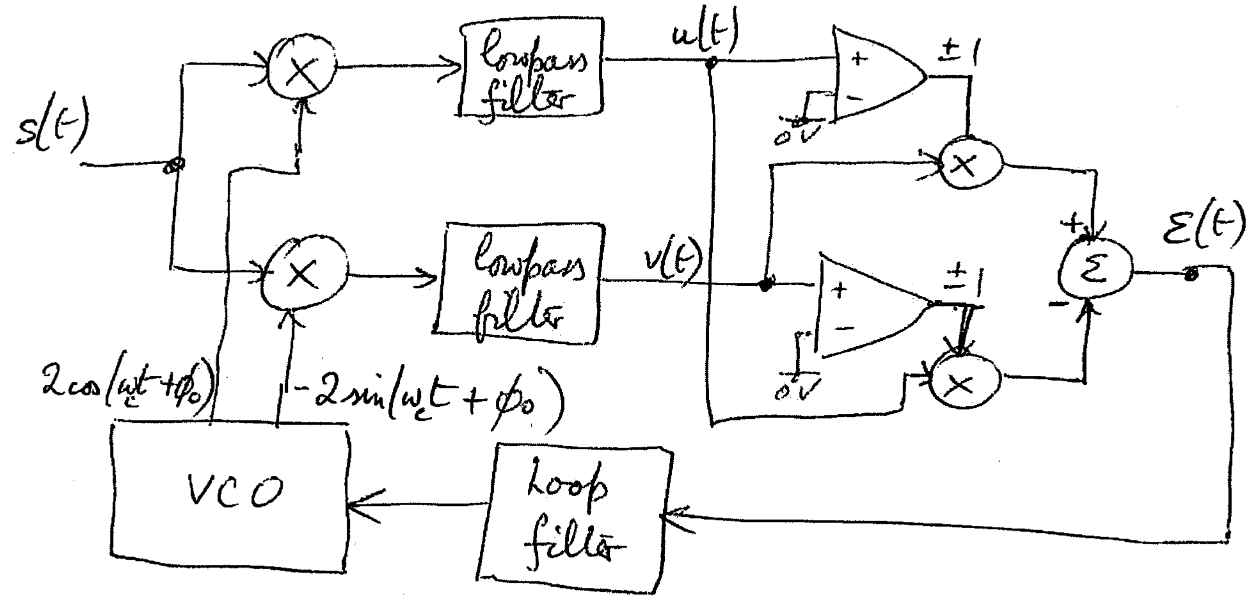
3(b) (cont) Block diagram:



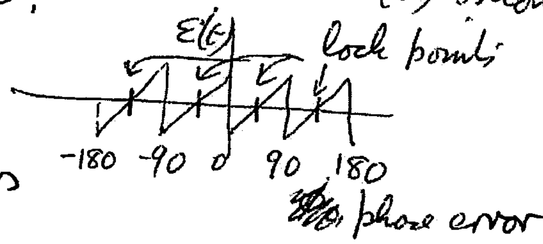
The $u(t)$ output of the quadrature demodulator can be used to replace the complex multiplier & Re[·] operator, when the input signal is a real modulated signal $s(t)$. Usually $g(t-kT_b)$ is constant over the bit period from $t = kT_b$ to $(k+1)T_b$ & can be incorporated in G. The phase shift ϕ_0 is applied to the local oscillator cos & sin waves. [8]

(c) For QPSK we need to add a second Matched Filter, Threshold Detector & Sampling Register to the $v(t)$ output of the Quadrature Demodulator, in order to demodulate the second stream of symbols that are modulated onto the quadrature carrier, as shown above. [3]

(d) A Phase-Locked Loop (PLL) for QPSK demodulation is typically as follows:



The loop error signal $\varepsilon(t)$ is designed to keep the VCO phase-locked to one of the four main phases of the input signal $s(t)$. If there is no phase shift on $s(t)$, then b_{2k} modulates the inphase component to be at $0°$ or $180°$, and b_{2k+1} modulates the quadrature component to be $\pm 90°$. Hence the resultant is at $\pm 45°$ or $\pm 135°$. In this case the phase ϕ_0 of the VCO must be $0°$. The error $\varepsilon(t)$ should have a characteristic that is



so that it locks with phase errors of $\pm 45°$ or $\pm 135°$.

[5]

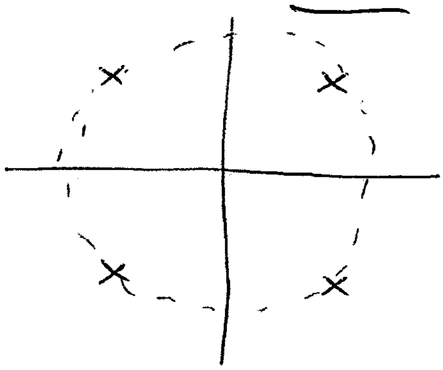
4 (a) $U = \dfrac{\text{Energy per bit}}{\text{Noise PSD}}$

U is a dimensionless quantity which specifies the signal-to-noise power ratio at the detector of a digital demodulation system. Different types of modulation result in different scaling coefficients. Modulation of the BPSK, QPSK, & QAM formats typically produce bit-error probabilities that are proportional to $Q(\sqrt{KU})$ where K is a constant dependant on the format & $Q(\;)$ is the Gaussian error integral function.

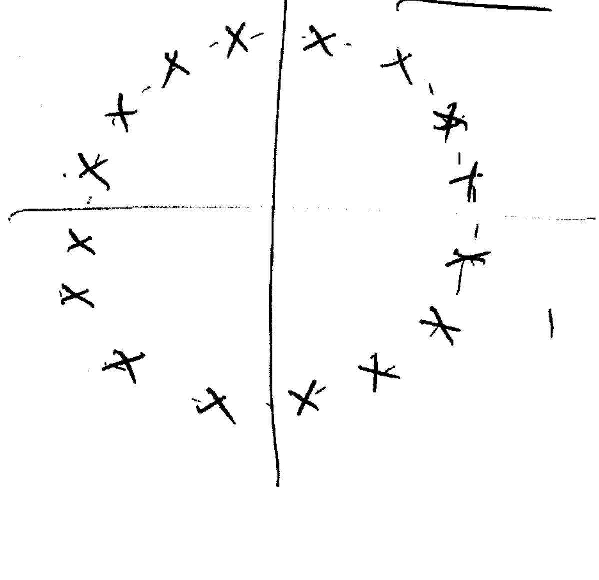
The power of the signal is (Energy per bit) $\times$ (bit rate), so if the noise PSD remains constant, the signal power needs to be proportional to the bit rate in order to maintain constant bit-error ~~rates~~ probability. [4]

(b)

QPSK

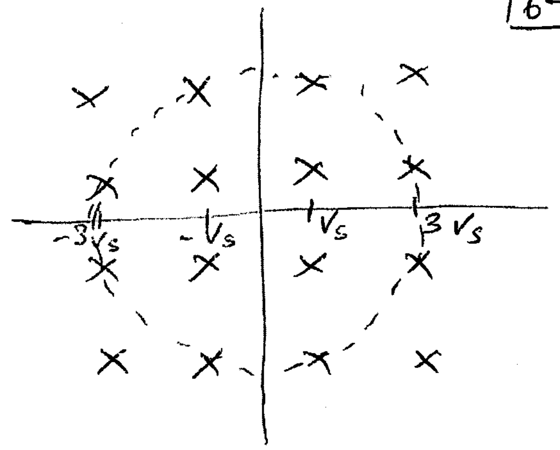


16-PSK



4 (b)

16-QAM



1

For QPSK, there are $m=2$ bits transmitted per symbol ($2^m = 4$ levels).

For 16-PSK & 16-QAM, there are $m = 4$ bits transmitted per symbol ($2^m = 16$).

Hence if the energy per bit is constant, the energy per symbol is doubled when m changes from 2 to 4 bits/sym. This will increase the rms amplitude of the 16-level constellations by $\sqrt{2}$ compared to QPSK (ie the 16-PSK circle is $\sqrt{2}$ times the diameter of the QPSK circle).

However it is clear that, despite this $\sqrt{2}$ increase, the constellation points for 16-PSK are much closer together than the QPSK points. Since the noise amplitude will be unchanged, the probability of symbol error for 16-PSK will be significantly greater due to closeness of the decision thresholds (half-way between the points). 1

4 (b) For 16-QAM, the situation is a bit less clear-cut. The rms amplitude of the 16-QAM constellation if the points are at odd multiples of V_s ~~[illegible]~~ is :

$$V_{rms}^2 = \frac{V_s^2}{4}\left(\cancel{4}(1^2+1^2)+(3^2+1^2)+(1^2+3^2)+(3^2+3^2)\right)$$

$$= \frac{V_s^2}{4}\,(2+10+10+18) = 10\,V_s^2$$

$\therefore\ V_{rms} = \sqrt{10}\,.\,V_s$ 1

This is the radius of the 8 points on the dashed circle shown,

For 16-QAM, distance between adjacent points $= 2V_s = \frac{2}{\sqrt{10}}V_{rms} = 0.6325\,V_{rms}$

For 16-PSK, " " " " $\approx V_{rms}\cdot\frac{2\pi}{16} = 0.3927\,V_{rms}$ · 1

Hence the 16-QAM will have better error probability than 16-PSK, because the constellation points are further apart for the same mean signal power. ~~F~~

(However 16-QAM will not have as low error probability as QPSK.) 1

The one advantage of 16-PSK (and QPSK) over 16-QAM is that it is a constant amplitude modulation scheme, so the signal detection thresholds are not affected by non-linear channels or amplitude fading. 1

[8]

4. c) Since the symbol rate for QPSK is twice that of 16-PSK or 16-QAM, the QPSK bandwidth will be ~~[illegible]~~ twice that of the other schemes, and three times that of ~~a~~ 64-QAM (6 bits per sym).

<u>BUT</u> 16-QAM & 64-QAM are more sensitive to noise and intersymbol interference than QPSK.

Hence for ~~DA~~ digital audio (DAB), where the bit rates are not too high (due to audio compression & the inherent lower bandwidth of audio) QPSK is selected in order to provide the most resilient modulation format to noise etc, to allow good mobile reception.

For digital video (DVB), bit rates are much higher, even when compression is used, & most TV receives used fixed directionally-selective antennae, so a more spectrally efficient modulation format is chosen (eg 64-QAM) to keep bandwidth usage to a minimum. The greater sensitivity to noise etc is tolerated in order to achieve this. (64-QAM needs ~8dB more signal power than QPSK for the same error rate.)

[4]

4.d) ~~To reduce the~~ Multiple transmission paths with different path delays cause adjacent symbols to interfere with each other (inter-symbol interference, ISI) if the path delay spread exceeds the symbol period or is comparable to it. Hence, for a given delay spread τ, we must ~~reduce~~ increase the symbol period to be $\gg \tau$. This is achieved by orthogonal frequency division multiplexing (OFDM).

OFDM splits the serial input data stream (at typ ~~...~~ 40 Mb/s for coded DVB) into ~~many~~ N parallel streams at $\frac{1}{N}$ of the data rate each. Each stream is modulated onto a separate subcarrier (using an inverse FFT process for efficiency) and the whole block of subcarriers is transmitted in approx the same bandwidth as the original serial signal would have required.

By using ~~the~~ guard bands between FFT blocks, orthogonality of the subcarriers can be maintained at the receiver despite path delay spreads up to the duration of the guard band. Hence the name OFDM. The FFT blocks are ~~periodically~~ cyclically extended into the guard band at the transmitter, so that for any path delay within the limits of the guard band, a pure sinusoid is received during the receiver analysis block period. [4]

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Engineering Triops Part 2A
Module 3F4. Data Transmission, May 2011 - Answers

1.

a)

b) (i) $|H_R(\omega)| = \sqrt{\frac{T}{\sqrt{N_0}}}\cos\left(\frac{\omega T}{2}\right)$

$|H_T(\omega)| = k\sqrt{T\sqrt{N_0}}\cos\left(\frac{\omega T}{2}\right)$

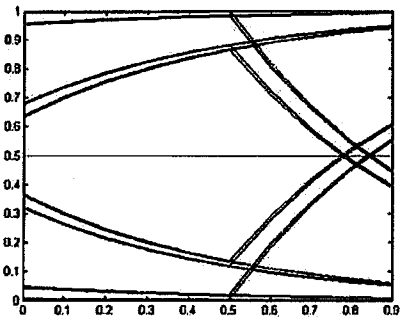
(ii) $\frac{3\pi^2 T}{32 N_0}$

2.

a)

b)

c) (i)



(ii) 2.01×10^{-3}

(iii) 0.5V, 1.31×10^{-4}

3.

a)

b)

c)

d)

4.

a)

b)

c)

d)

I. J. Wassell

May 2011