DESIGN AND DEVELOPMENT OF DATA RECOVERY SYSTEM FOR FADED WIRELESS COMMUNICATION CHANNEL

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ABSTRACT

Wireless communication channel is being used as primary means of communication nowadays. Data is DPSK modulated and hamming encoded before being given as input to the Rayleigh Faded Wireless Communication channel. This paper mainly focuses on the data re-generation techniques from phase modulated data at the source side. The generated data is made to pass through the channel by undergoing various transmission impairments. To accurately regenerate the data at the receiver side equalizers are recommended for usage. Decision feedback equalizer is one such equalizer that reproduces the data accurately. It takes care of all types of the noise components and suppresses the AWGN also. It counters the Doppler Effect and ISI most accurately.

Keywords: Feedback Equalizer, Doppler correction, I channel and Q channel.

I DATA GENERATION

The input to the system is the message of 'L' number of characters. The model converts the given text into ASCII code and supplies the bit stream to DPSK encoder[3].

The general representation for a set of M-ary phase signaling waveform is[6][3]

$$S(t) = \{U(t) \exp j\, (2\Pi f_c\, t + 2\Pi\, (m-1) / M + \theta)\}$$

$$0 \le t \le T,$$
$$m = 1,2,3,4$$
$$\theta = \text{initial phase.}$$

Where U(t) is a rectangular pulse with amplitude 'A'

$$S(t) = A \cos [2\Pi f_c\, t + 2\Pi\, (m-1) / M + \theta]$$

$$= A_{\text{Channel Simulation}} \cos [2\Pi f_c\, t] - A_{em} \sin[2\Pi f_c\, t]$$

Where $A_{\text{Channel Simulation}} = \cos [2\Pi\, (m-1) / M + \theta]$

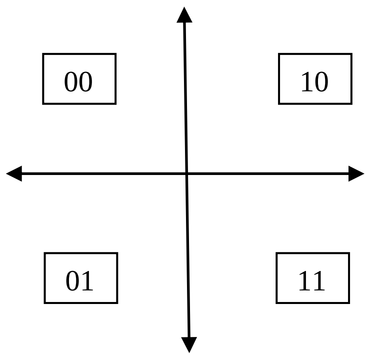
$$A_{em} = \sin [2\Pi\, (m-1) / M + \theta]$$

If $\theta = \Pi/4$ and M = 4, then $A_{\text{Channel Simulation}} = A_{em} = \pm A/2$.

The mapping or assignment of information bits to the M possible phases more commonly done using the logic described here under as shown in Fig (1).

In a four phase PSK, sets of two successive bits are mapped on to the four possible phases. When a pair of bits is encoded, say 0 1, the phase corresponding to this combination i.e., 5Π/4, added to the phase shift corresponding to the previous bit interval, say 7Π/4, to give the phase shift the present interval[6].

Thus while in the previous bit interval a sinusoid of signaling frequency f_c with phase shift 7Π/4 was transmitted, in the present interval the sinusoid is transmitted with a phase shift of 7Π/4 + 5Π/4 = Π.



Bits	**Code**
00	011
01	101
10	001
11	111

Bits	**Phase**
00	3/ 4π
01	5/4 π
10	1/ 4 π
11	7/4 π

Phase	**Code**
0 π	000
1/ 4 π	001
2/4 π	010
3/ 4 π	011
4/4 π	100
5/4 π	101
6/4 π	110
7/4 π	111

Fig (1) Signal Constellation for 4 ary PSK

When θ > 0 is used, the signaling phase is shifted in every signaling interval, even when a long string of zeroes occur in the information. This results in a signal spectrum with a width that approximately equal to 1/T. (T is the signaling interval). The spectral components above and below the carrier are used in maintaining synchronization at the receiver. Hence their presence in the received signal is important. Thus a non zero value of 'θ' (θ = π/4) is used in such a case. With equal to four and θ equal to π /4, eight possible phase shifts exist.

The sequence of pulses $\{\sum_k a_k\}$ or $\sum_i S_i\, \delta(t - iT)$ from DPSK is fed to the radio equipment transmitter filter 'G'. It has an impulse response of g(t) and a transfer function of G(f). The output this filter is real-valued and is given by

$$S(t) = R\left[\sum_i S_i\, \delta(t - iT)\right] * g(t) \quad \text{(* represents Convolution)}$$

$$= \sum_i S_i\, \delta(t - iT) * g(t) + \sum_i S_i^*\, \delta(t - iT) * g(t)$$

$$= \sum_i S_i\, g(t - iT) + \sum_i S_i^*\, g(t - iT)$$

II DATA TRANSMISSION

When S(t) is fed into a single Rayleigh fading HF channel the output would be [3][7][4]

$$-x(t) = S'(t)q_1(t) + S^{\wedge}(t)q_2(t)$$

where q_1 (t) and q_2 (t) are random process.

The random processes have been generated by filtering the zero mean white Gaussian noise signal through a Besel filter. The frequency and impulse response of Bessel filter approach Gaussian when the order of the filter is sufficiently large. For the three sky waves, the model has, which requires six random processes $q_1(t)$ to $q_6(t)$. The variance of all six variables is equal to 0.167. It ensures that the total variance of the three sky wave channel is unity. Each of the values of $q_i(t)$ generated from an independent source so that they will be uncorrelated [4].

$$S^{\wedge}(t) = S'(t) * f(t)$$

Where f(t) is the impulse response of a Hilbert Transform filter whose Fourier transform given by [5]

$$F(f) = j \quad f < 0$$
$$= 0 \quad f = 0$$
$$= -j \quad f > 0$$

$$S^{\wedge}{}'(T) = \{ \textstyle\sum_i S_i\, g(t - iT) + \sum_i S_i^*\, g(t - iT) \} * f(t)$$
$$= \textstyle\sum_i S_i\, (g(t - iT) * f(t)) + \sum_i S_i^*\, (g(t - iT) * f(t))$$

Where S_i^* is complex conjugate of S_i

Let us consider that the modeled channel has three independent Rayleigh-fading sky waves. The relative delay of the two sky waves from direct path be 'τ' seconds and 'τl' seconds respectively. The output from the channel is given by [8]:

$$x(t) = [S'(t)\, q_1(t) + S^{\wedge}{}'(t)\, q_2(t)] + [S'(t - \tau)\, q_3(t) + S'(t- \tau)\, q_4(t)]$$
$$+ [S'(t-\tau l)\, q_5(t) + S'(t- \tau l)\, q_6(t)]$$

While passing through the channel, the signal is modified as a result of random noise n(t) added to it. The noisy signal x'(t) is then passed through receiver filter C(t). The receiver filter output is given by

$$y(t)= x(t) * C(t) + n(t) * C(t)$$

The output of the matched filter d(t) is given to the demodulator where the signal splits into inphase (I) and quadrature (Q) channels.

I channel input is given by

$$y'(t) = \{[x(t) * C(t) * d(t)] \exp(-j2\Pi fct)\}$$
$$+ \{[n(t) * C(t) * d(t)] \exp(-j\, 2\Pi f_c\, t)\}$$

Where n(t) = noise produced at the output of receiver filter due to additive noise AWGN at the receiver input.

This can be written as

$$y'(t) = \{ \mu \textstyle\sum_I a_i\, p(t-kT) n(t) \} \exp(-j2\Pi f_c t)$$

Where μ = Scaling factor,

p(t) = Defined pulse,

p(0) = 1,

$\mu p(t)$= x(t) * C(t) * d(t);

$$y'(t_i) = \mu\, a_i + \mu \sum_{\substack{k \neq i \\ k=-\infty}}^{\infty} a_k p[(i-k)T] + n(t_i)$$

Where 1st term = Contribution of the Ith transmitted bit, 2 nd term = Residual effect of all other transmitted bits on the Ith bit,

3 rd term = Noise sample at t.

The residual effect due to occurrence of pulses before and after the sampling instant t_i called ISI. In the absence of ISI and noise,

$$y(t_i) = \mu a_i$$

III DATA RECOVERY

In the absence of noise, in the ith signaling interval, the received signal can be simply represented as [9]

$R_i(t) = \sin(2 \Pi f_c t + \theta + \phi_t)$

Where fc = Carrier frequency having a phase shift of $(\theta + \phi_t)$

ϕ_t = Information bearing

θ = Unknown phase shift which represents the path delay and relative phase difference between transmitter and receiver oscillators.

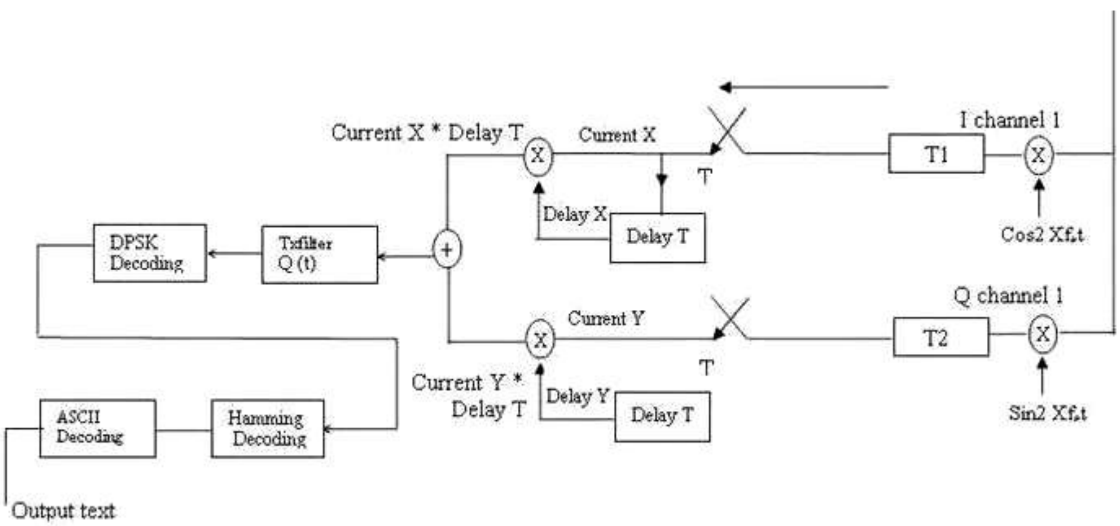


Fig (2) Model of Data Receiving System over Radio Link

The in-phase (I) channel of demodulator gives

$$Xc = \int_{T_1}^{T_2} S_i(t) \cos(2\Pi f_c t)\, dt$$

$$= \int 1/2 \left(\sin(4\Pi f_c t + \theta + \phi_t) + \sin(\theta + \phi_t)\right) dt$$

$$= l/2 \sin(\theta + \phi_i)$$

Where T_1 -T_2 corresponds to a time period in which f_c has an integral number of cycles.

$X_p = X_{i-1} = Xc$ in the previous signaling interval.

$$= 1/2 \sin(\theta + \phi_{i-1})$$

assuming path delay is constant for two Successive signaling intervals.

Similarly the quadrature phase (Q) channel of demodulator will give

$$Yc = \int_{T_1}^{T_2} S_i(t) \sin(2\Pi f_c t)\, dt$$

$$= \int 1/2 (\cos(\theta + \phi_t) - \cos(4\Pi f_c t + \theta + \phi_t))\, dt$$

$$= l/2 \cos(\theta + \phi_i))$$

$$Y_p = Y_{i-1} = Yc \text{ in the previous signaling interval.}$$

$$= 1/2 \cos(\theta + \phi_{I-1})$$

The final outputs of the system are R and Q signals.

$$R = X_c * X_p + Y_c * Y_p$$

$$= k \sin(\theta + \phi_i) \sin(\theta + \phi_{i-1}) + k \cos(\theta + \phi_i) \cos(\theta + \phi_{i-1})$$

and

$$Q = X_c * Y_p - Y_c * X_p$$

$$= k \sin(\theta + \phi_i) \cos(\theta + \phi_{i-1}) - k \cos(\theta + \phi_i) \cos(\theta + \phi_{i-1})$$

The decision is now made on the basis of the signs of 'R' and 'Q' signals as given under.

$R < 0$ & $Q < 0$ 01 ; $R < 0$ & $Q > 0$ 00 ;

$R > 0$ & $Q < 0$ 11 ; $R > 0$ & $Q > 0$ 10 ;

The period of integration $T_2 - T_1$ makes it possible to avoid the use of band-pass filter separate out noise (if any) from received signal.

During data demodulation, the carrier f_c generated at receiver should not run freely. It must be locked to a fixed phase at the beginning of every signaling interval. This is necessary to avoid a differential phase shift in the locally generated carrier, multiplying the incoming phase in successive signaling intervals. This differential phase shift would other wise add to the information bearing phase shift ($\phi_i - \phi_{i-1}$) and cause erroneous decisions to be taken at the demodulator.

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