A COHERENT DIGITAL DEMODULATION TECHNIQUE FOR CONTINUOUS PHASE MODULATION (CPM)

by

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Abstract

This paper presents a novel method of recovering both the clock and carrier from the received signal at baseband for the class of continuous phase modulated (CPM) signals, Minimum Shift Keying (MSK), Gaussian Minimum Shift Keying (GMSK), Tamed Frequency Modulation (TFM) and Generalised Tamed Frequency Modulation (GTFM). The technique adopted for this particular receiver uses zero crossings of the phase of the received signal as the only source of information about both the clock and carrier, Computer simulations are presented for difference digital modulation schemes in the presence of Additive White Gaussian Noise (AWGN) with offsets in clock phase and carrier phase. Bit error rates for different modulation schemes are also presented.

1.0 Introduction

One class of digital signals that is particularly suited to radio is Minimum Shift Keying (MSK) and its derivatives Gaussian Minimum Shift Keying (GMSK), Tamed Frequency Modulation (TFM) and Generalised Tamed Frequency Modulation (GTFM). The possibility of fully integrated radio receivers has renewed interest in direct demodulation (zero IF) techniques and it may be observed that the coherent demodulation for this class of digital signals is almost identical in layout to a direct demodulation radio receiver. Hence possibility arises of combining the two functions.

This paper presents a combined receivers and demodulator where the modulated RF signal is shifted down to baseband by a pair of quadrature mixers and then lowpass filtered. Since the lowpass filters provide the selectivity of the receiver both the carrier and clock recovery are performed down stream of these filters. Since both the clock and carrier signals are recovered from the baseband signal, this particular scheme is different from many versions of the conventional coherent demodulator where the carrier recovery is achieved by a phase locked loop (PLL) at RF or at a convenient non zero IF. Since both the clock and carrier signals are recovered from the zero crossings of the hard limited baseband signals, the technique employed allows simultaneous acquisition of the clock and carrier and hence is suitable for systems requiring fast acquisition.

Following is a brief summary of the paper. In Section 2.0 a detailed description of the novel demodulation technique is presented. Here the resultant effect of offsets in carrier phase and clock phase on the phase of the received signal in the presence of additive white Gaussian noise (AWGN) is discussed in detail. Section

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3.0 is a presentation of the simulation results. The performance of the demodulation technique for MSK, GMSK, TFM and GTFM both in the absence and presence of additive white Gaussian noise, and the bit error rate performances of MSK, GMSK, TFM and GTFM are presented in this respect. In Section 4.0, conclusions are drawn about the performance of the demodulation technique based on the other conventional coherent domodulation methods.

2.0 The demodulation technique and its performance

Fig. 2.1 shows the form of a combined receiver and demodulator used in this work (5,6), where both the clock and carrier signals are recovered solely from the information contained in the timing of the zero crossings in the hard limited outputs of the two quadrature channels (3).

The technique will be explained with reference to Fig. 2.2, in which the phase trajectories shown represent, noise free MSK with no bandwidth restrictions (the curve with sharp angles) and TFM, GTFM, GMSK (the smooth curve). The trajectory represents the phase of the modulated RF signal relative to its carrier, and the phase of the signal is transferred directly through the two quadrature mixers, subject only to two constant shifts separated by 90 degrees, And in the diagram the phase axis is labeled with the equivalent phases at the outputs of the inphase (I) and quadrature (Q) channels.

The zero crossings in the I and Q channels correspond to the trajectory crossing the horizontal lines which are spaced at 90 degrees of phase. The solid lines represent zero and 180 degree phases in the Q channel and the dashed lines the zero and 180 degree phases in the I channel. The solid vertical lines (dashed vertical lines) represent the instants at which the Q channel (I channel) is expected to have a transition and the I channel (Q channel) to be at the center of its eye.

An error in the carrier phase is reflected as a vertical shift of the trajectory and, an error in the clock phase is reflected as a horizontal shift. The general demodulation technique will be described with reference to Fig. 2.3, which shows errors in both the clock and carrier phases, where the clock is running early by x and the carrier oscillator phase early by y.

The error observed in the times of zero crossings depends on the slope of the trajectory, that is on whether the instantaneous frequency is higher or lower than the carrier. Therefore in principle, on a noise free channel, with an unlimited bandwidth MSK signal, only one high frequency transition and one low frequency transition are required to set the phase of both the clock and carrier oscillators. This can be done easily in two simple steps in the following manner.

From Fig. 2.3, by simple geometric properties, we can write

$$EH = x - y \qquad (2.1)$$

$$EL = x + y \qquad (2.2)$$

By solving (2.1) and (2.2), we have

$$x = (EL + EH)/2 \qquad (2.3)$$

$$y = (EL - EH)/2 \qquad (2.4)$$

If the first transition is due to a low frequency, then we measure EL = (x + y). Apply a phase correction of (x+y)/2 to both the clock and carrier oscillators. This cause any subsequent low frequency transition error to become zero, but next high frequency transition error will become (x-y). Then a correction of (x-y)/2 is applied to the clock and -(x-y)/2 to the carrier, causing zero residual errors in both.

In practice the signal is always

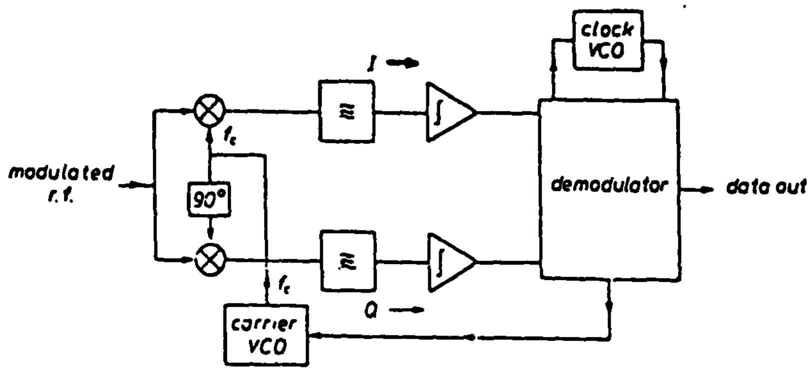

Fig. 2.1 Combined receiver and demodulator

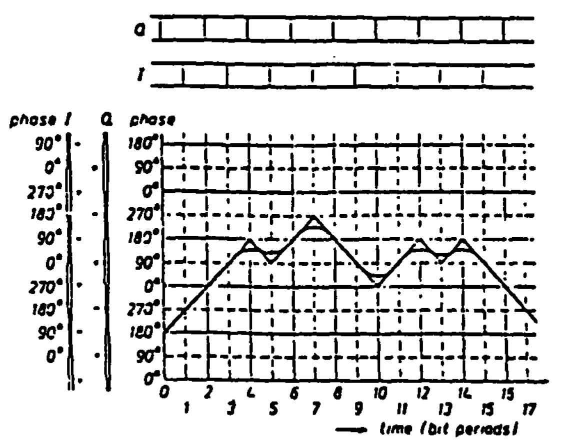

Fig. 2.2 Phase trajectories of CPM signals

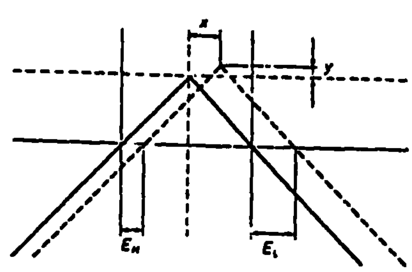

Fig. 2.3 Combined effect of clock and carrier
phase errors

bandlimited and the phase trajectory is not a set of straight lines and sharp angles. Nevertheless, a partial correction algorithm will converge the process on the desired position. The optimum strategy will be dependent on the requirements on error performance and acquisition time, and also on the relative stabilities of the clock and carrier oscillators. The corrections to the phase of the clock and carrier oscillators can be derived from the following rules.

a) CLOCK

If a transition is early, advance the clock and if it is late, retard the clock.

b) CARRIER

Here the same rule as above or its inverse applies, according to whether the transition was due to a high or a low frequency.

Hence the rule is,

transition	frequency	action
late	low	advance
late	high	retard
early	low	retard
early	high	advance

Whether a transition corresponds to a high, or a low frequency can be determined by observing the sign of the other channel at the moment of the occurrence of the transition. That is, if they have the same sign, it corresponds to a high frequency transition and if they have opposite signs, it corresponds to a low frequency transition. In many cases the carrier phase will need more frequent correction than the clock and a relative clock oscillator will not be unduly disturbed by carrier phase fluctuations. The phase modulated signal in the presence of AWGN, is represented by

$$s(t) = A\cos[2\pi f_c t + \phi(t)] + N(t) \qquad (2.5)$$

where, N(t) is stationary, Gaussian, random variable with a distribution $(0, \sigma^2)$, and A is a constant. N(t) can be expressed in its quadrature representation as

$$N(t) = n_I(t)\cos(2\pi f_c t) - n_Q(t)\sin(2\pi f_c t) \qquad (2.6)$$

where, $n_I(t)$ and $n_Q(t)$ are two Gaussian, statistically independent, stationary random variables with identical distributions $(0, \sigma^2)$.

By defining the new variables, X(t) and Y(t), the effect of noise could be incorporated in the phase of the transmitted signal in the following manner (7).

$$X(t) = n_I(t)\cos[\phi(t)] + n_Q(t)\sin[\phi(t)] \qquad (2.7)$$

$$Y(t) = n_Q(t)\cos[\phi(t)] - n_I(t)\sin[\phi(t)] \qquad (2.8)$$

substituting in (2.5) and rearranging terms, s(t) could be rewritten as

$$s(t) = R(t)\cos[2\pi f_c t + \phi(t) + \alpha(t)] \qquad (2.9)$$

Where, $R(t) = [(A+X(t))^2 + Y(t)^2]^{1/2}$

$$\alpha(t) = \tan^{-1}[Y(t)/(A+X(t))]$$

A block diagram for the combined receiver and demodulator for MSK is shown in Fig. 2.4.

The received phase modulated signal s(t), which is given by (2.9) is multiplied by the respective inphase and quadrature carrier components, followed by the respective shaping waveforms (4). The multiplier-integrator combination constitutes a correlation detector, which is equivalent to matched filtering.

Since the correlation detection is equivalent to matched filtering, the output of each correlator (input to the sampler) is equal to the low frequency component contained in the inputs to the quadrature $[r_Q(t)]$ and inphase $[r_I(t)]$ arms,

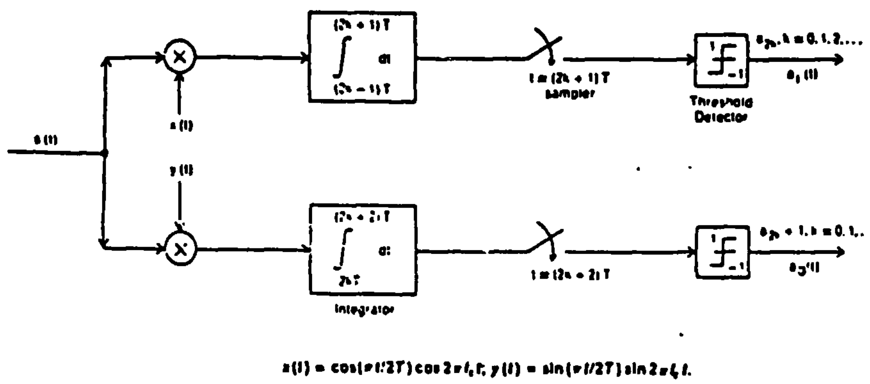

Fig. 2.4 Block diagram of the combined demodulator and receiver for MSK

given by

$$r_I(t) = 0.5R(t)[\cos\theta(t)\cos(\pi t/2T) + \cos(\theta(t))\cos(\pi t/2T)\cos(4\pi f_c t) - (\sin\theta(t)\cos(\pi t/2T)\sin(4\pi f_c t))] \quad (2.10)$$

$$r_Q(t) = 0.5R(t)[-\sin\theta(t)\sin(\pi t/2T) + \cos(\theta(t)\sin(\pi t/2T)\sin(4\pi f_c t) + \sin\theta(t)\sin(\pi t/2T)\cos(4\pi f_c t)] \quad (2.11)$$

where, $\theta(t) = \phi(t)+\alpha(t)$

Therefore the inputs to the samplers in the inphase and quadrature arms are given by $0.5R(t)\cos[\theta(t)]\cos(\pi t/2T)$ and $(-0.5)R(t)\sin[\theta(t)]\sin(\pi t/2T)$ respectively. Sampling is performed at instants $(2k+1)T$ and $(2k+2)T$ respectively at the inphase and quadrature arms of the demodulator and the polarity of sampler outputs determines the value of $a_I(t)$ [$a_Q(t)$] in the inphase channel [quadrature channel]. Bit error rates are calculated by using a random bit sequence at the MSK modulator, and comparing this with the output of the demodulator discussed above. Before the bit error calculations are performed, clock and carrier synchronization is achieved by, using the correction algorithm presented earlier in the chapter.

Receivers for GMSK, TFM and GTFM are very similar to MSK, the only difference being the receiving filters R(f) following the inphase and quadrature multipliers. Other than that the operation of the combined receiver and demodulator for GMSK, TFM and GTFM is identical to that of MSK. The general block diagram of the combined receiver and demodulator for GMSK, TFM and GTFM is shown in Fig. 2.5.

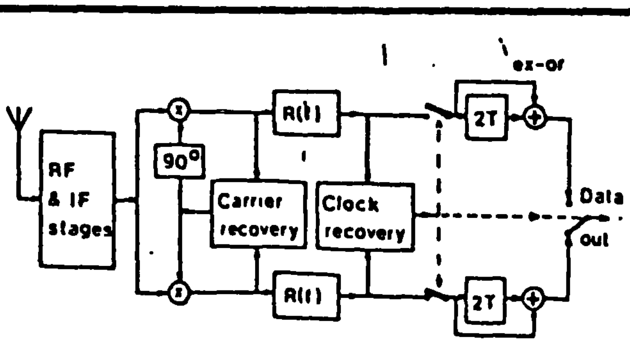

Fig. 2.5 Block diagram of the combined receiver and demodulator for GMSK, TFM and CTFM

The receiving filter for TFM is given by (2)

$$R(f) = A[\sin(4\pi fT)/4\pi fT \quad 1-(4fT)2\ ] \tag{2.12}$$

where A is an arbitrary constant.

The receiving filter for GTFM consists of a Gaussian lowpass filter, D(f) in cascade with a 3-tap transversal filter, Y(f) given by (1)

$$R(f) = D(f).Y(f) \tag{2.13}$$

$$R(f) = D(f)\ [1+2e\ \cos(4\pi fT)] \tag{2.14}$$

where D(f) will typically have a normalised bandwidth between 0.2 and 0.24, and Y(f) has tap delays adjusted (by varying the outer tap coefficient e) to reduce intersymbol interference caused by severe band-limitation of D(f).

3.0 Simulation results

A signal to noise ratio of 10 dB was chosen for calculation purposes, and a bit rate of 64 Kbit/s and an interpolation factor (number of samples for bit period) of 16 were used for all calculation purposes throughout the simulation. The incoming data sequence

ak = -1. -1, -1, -1, 1, -1, 1, -1,
1, 1, -1, -1, 1, 1, 1,
-1, -1, 1, 1

was used to generate the phase trajectories for different signalling schemes. An initial offset of 0.25 radians and 0.25T (bit period) were assumed for carrier phase and clock phase respectively, for all signalling schemes. A roll-off factor (r) of 0.5 and a tap-coefficient (B) of 0.6 were assumed in the generation of the GTFM signal.

To ensure reliability of the simulation results, the bit error rates are computed from a minimum of alteast 100 errors for each value of signal to noise ratio. The bit error rate performances of MSK, GMSK, TFM and GTFM are compared with optimum PSK in Fig. 3.1.

From the bit error rate probabilities it could be seen that the best performance is obtained for MSK, in comparison to optimum PSK, and is only 0.6 dB below optimum PSK at a bit error rate of 10-4. Bit error rate performances of other signalling schemes are in the descending order GTFM, GMSK and TFM. At a bit error rate of 10-4 GTFM and GMSK are respectively 1.2 dB and 1.3 dB inferior to optimum PSK and TFM is 2 dB below optimum PSK at a bit error rate of 10-4. It is also evidenced that the bit error rate performances of GTFM and GMSK are very similar and this could be attributed to the almost identical phase trajectories of GTFM and GMSK.

4.0 Conclusions

The coherent digital demodulation technique presented in this paper was shown to be a fast and fairly

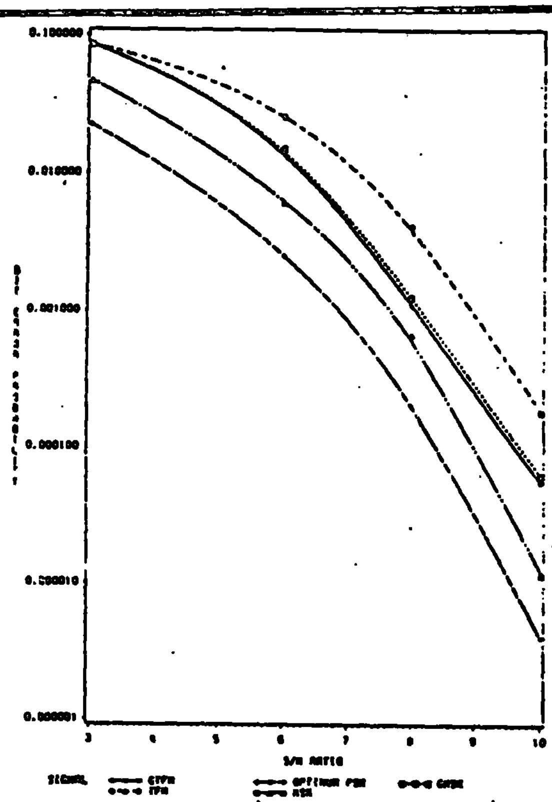

Fig. 3.1 Bit error Rate Perforamnce of MSK, GMSK, TFM and GTFM
(r = 0.3, B = 0.6)

accurate method of recovering both the clock and carrier for the class of continuous phase modulated (CPM) signals MSK, GMSK, TFM and GTFM.

One important advantage of this technique is the ease with which it could be implemented, due to the fact that, the demodulator makes use of only the zero crossing information contained in the hard limited baseband signals, for synchronization of both the clock and carrier. Since the technique adopted allows simultaneous acquisition of the clock and carrier, the suggested demodulation technique is suitable for systems requiring fast acquisition.

The algorithm for the case of offsets in carrier phase and clock phase has shown promising results, both in the absence and presence of additive white Gaussian noise, as evidenced from the simulation results. If the carrier phase offset obtained from the first detected low and high frequency transitions is not sufficiently accurate, the algorithm applies a series of partial corrections on the phase offset until a fairly accurate result is obtained. This was necessary, especially in the cases of TFM, GTFM and GMSK.

Bit error rates obtained for MSK seemed to be closer to theoretical limits, especially at higher signal to noise ratios. This could be attributed to the highly accurate synchronization obtained by the demodulation algorithm for both the carrier and clock. A major factor that drastically affects the bit error rate performance, is the clock offset, which is the main source of intersymbol interference. Since the algorithm has achieved almost perfect clock synchronization, as it is evidenced from the simulation results, the observed bit error probabilities are extremely low. The simulated performance for MSK at a bit error rate of 10-2 is 2.6 dB superior to the performance obtained in (3) for the practical realization of the demodulation algorithm, with perfect synchronization of clock and carrier. This impairment in bit error rate performance could be attributed mainly to the practical, non-optimum filters used in the I and Q channels. In this thesis optimum filters were assumed for I and Q channels in bit error rate calculations.

Bit error rate performance of TFM and GTFM closely follow the simulated results for coherent detection of TFM and GTFM with perfect carrier and clock synchronization, given in (1). From this it is evidenced that the demodulation algorithm has only about 0.2 - 0.4 dB degradation on the bit error rate performance of TFM and GTFM.

5.0 References

1. K.S. Chung, "Generalised Tamed Frequency Modulation and its Application for Mobile Radio Communs", IEEE Trans. Vehicular Tech., Vol. VT 33, No. 3, pp. 103-113, August 1984.

2. K. Murota and K. Hirade, "GMSK Modulation for Digital Mobile Radio Telephony", IEEE Trans.Commun., Vol COM 29, No. 7, pp. 1044-1050, July 1981.

3. R.J. Murray and W. Gibson, "A Coherent Digital Demodulator for MSK and related Modulation Schemes", Philips J. Res. 39, Nos. 1/2, 1984.

4. S. Pasupathy, "Minimum Shift Keying, A Spectrally Efficient Modulation", IEEE Commun. Mag., Vol. 17, pp. 14-22, July 1979.

5. P. Satarasinghe, "A Coherent Digital Demodulation Technique for Continuous Phase Modulation Schemes", Masters Research Thesis, School of E.E.C.S., University of Oklahoma, Februrary 1988.

6. P. Satarasinghe and J.R. Cruz, "A Coherent Digital Demodulation

Technique for Continuous Phase Modulation", Technical Report ++15-003, School of E.E.C.S. University of Oklahoma, February 1988.

7. H. Schachter, D.L. Schilling and J.J. Refi, "Output Signal to Noise Ratio of an FM Discriminator with non ideal Limiting", IEEE Trans. Commun. Tech., pp. 673-676, October 1970.