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A New Coherent Demodulation Technique for Land-Mobile Satellite Communications

Shousei Yoshida, Hideho Tomita
C&C Systems Research Labs., NEC Corp.
1-1, Miyazaki 4-chome, Miyamae-ku, Kawasaki-city, 213, JAPAN
PHONE +81-44-856-2255 FAX +81-44-856-2061

ABSTRACT

This paper describes an advanced coherent demodulation technique for land-mobile satellite (LMS) communications. The proposed technique features a combined narrow/wide band dual open loop carrier phase estimator, which is effectively able to compensate the fast carrier phase fluctuation by fading without sacrificing a phase slip rate. Also included is the realization of quick carrier and clock re-acquisition after shadowing by taking open loop structure. Its bit error rate (BER) performance is superior to that of existing detection schemes, showing a BER of 1×10^{-2} at 6.3 dB E_b/N_0 over the Rician channel with 10 dB C/M and 200 Hz (1/16 modulation rate) fading pitch f_d for QPSK. The proposed scheme consists of a fast response carrier recovery and a quick bit timing recovery with an interporation. An experimental terminal modem has been developed to evaluate its performance at fading conditions. The results are quite satisfactory, giving prospects for future LMS applications.

1. INTRODUCTION

Satellite based land-mobile communications have received increasing attention over recent few years because of its extended service to the rural area, where terrestrial mobile systems cannot be incorporated from an economical viewpoint. In LMS systems, the use of power and spectrally efficient transmission scheme is intensely requested to overcome the large space loss, and to conform the limited available frequency band. To meet such requirements, coherent detection combined with forward error correction (FEC) is in common practice. However, in the LMS environment, there exist two serious degrading factors for coherent detection. One is multipath fading, which ranges to 100- 200 Hz at maximum. Since a 4800 bps data rate is sufficient for a current digital voice transmission technique, good phase coherence cannot be counted on over long bit period, thus it results in serious BER degradation. The other is frequent received signal interruption by obstacles (i.e. shadowing). A phase lock loop (PLL) used conventionally in carrier and clock synchronization is very vulnerable to such environments, and often loses its track, taking a lot of time for the recovery. Although the phase tracking speed and/or recovery time can be improved by making loop bandwidth wide, it induces frequent phase slips as a penalty. It is quite difficult to find a good compromise between these two conflicting factors. Authors propose a new coherent demodulation technique, which is almost able to remove PLL's deficiency.

2. NEWLY PROPOSED COHERENT DEMODULATION SCHEME

2.1 Carrier Recovery by Dual Open Loop Phase Estimator

In order to resolve aforementioned problems, authors propose a new carrier phase estimator, which consists of two different bandwidth estimators to compensate the fast phase fluctuation by fading without increasing phase slips. An LMS channel is modeled by the mixture of a strong direct-path signal component and a scattered weak multipath signal component, that is Rician. Moreover the received signal is contaminated with thermal noise, as depicted in Fig. 1.

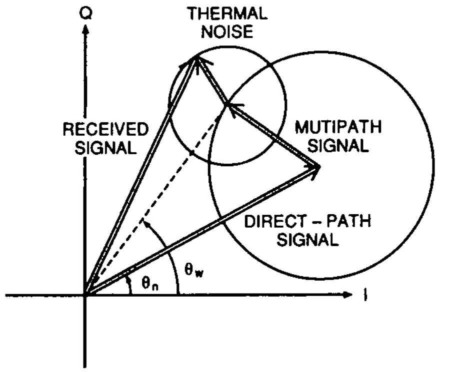

Fig. 1 Received signal vector over LMS channel

Figure 2 shows a block diagram of the proposed carrier recovery by a dual open loop phase estimator. This block diagram features open loop architecture used throughout to make free from acquisition problems. Basic open loop phase estimator's performance is previously analyzed in [1],[2]. A long term frequency offset is assumed to be corrected by an automatic frequency control (AFC) circuit. As a first operation, phase modulation is removed by an Mth-law operator, after that, the signal is routed both to narrow and wide band phase estimators. The narrow band estimator's bandwidth is designed around some tens Hz so as not to induce phase slips by thermal noise, and the estimator's filter is constructed by a finite impulse response (FIR) filter because of its well-defined delay time. As a result, almost all the multipath components as well as thermal noise are smoothed out, because the filter's bandwidth is much narrower than assumed maximum fading pitch 200Hz, and only a direct-path component is derived. A narrow band estimator's output $\theta_n(n)$, n as an integer, is expressed by

$$\begin{aligned}\theta_n(n) &= \arg[\sum_{k=n-(N1-1)/2}^{n+(N1-1)/2} C_n(n-k)\mathbf{S}_M(k)] \\ &= \arg[\frac{1}{N_1}\sum_{k=n-(N1-1)/2}^{n+(N1-1)/2} \mathbf{S}_M(k)] \\ &\qquad (C_n(i) = 1/N_1 \quad \text{for} \quad {}^{\forall}i), \qquad (1)\end{aligned}$$

where $\mathbf{S}_M(n)$ is a modulation removed signal, and $C_n(i)$ and N_1 are tap coefficients and tap number of the narrow band estimator respectively.

The fast phase fluctuation due to the multipath is estimated by the wide band estimator. The wide band estimator takes the same configuration as the narrow band estimator except its bandwidth. A wide band estimator's output $\theta_w(n)$ is also expressed by

$$\begin{aligned}\theta_w(n) &= \arg[\sum_{k=n-(N2-1)/2}^{n+(N2-1)/2} C_w(n-k)\mathbf{S}_M(k)] \\ &= \arg[\frac{1}{N_2}\sum_{k=n-(N2-1)/2}^{n+(N2-1)/2} \mathbf{S}_M(k)] \\ &\qquad (C_w(i) = 1/N_2 \quad \text{for} \quad {}^{\forall}i), \qquad (2)\end{aligned}$$

where $C_w(i)$ and N_2 are tap coefficients and tap number of the wide band estimator respectively. An impulse response of the respective estimators is designed rectangular to reduce a computational load.

A carrier phase is estimated by combining the narrow and wide band estimator's outputs so that the wide band estimator's output cannot induce carrier phase slips. First, the narrow band estimator's output is subtracted from the wide band estimator's output, and the result is divided by modulation phase M, thus a fast phase component scattered by the multipath is obtained within $\pm\pi/M$ radian, for example $\pm\pi/4$ radian for QPSK. The narrow band estimator's output is also divided by M, and then its phase range is extended to $\pm\pi$ radian by quadrant correction. A recovered carrier phase is obtained by adding the estimated fast phase component to the quadrant corrected slow changing phase. Thus, a recovered carrier phase $\theta_r(n)$ is calculated by using the following equation,

$$\begin{aligned}\theta_r(n) = \bmod\{&\frac{1}{M}\theta_n(n) + \frac{2\pi}{M}i(n) \\ &+ \frac{1}{M}\bmod[\theta_w(n) - \theta_n(n), 2\pi], 2\pi\}, \quad (3)\end{aligned}$$

where $i(n)$ denotes $\bmod\{i(n-1) - \mathrm{sign}[\theta_n(n) - \theta_n(n-1), 1], M\}$ for $|\ \theta_n(n) - \theta_n(n-1)\ | > \pi/M$ and $i(n-1)$ otherwise. In this carrier recovery scheme, a carrier phase slip rate depends only on the narrow band estimator's performance. Finally, the recovered phase is converted to a complex form, and a received signal is coherently detected by the complex carrier. Delay time T_1 and

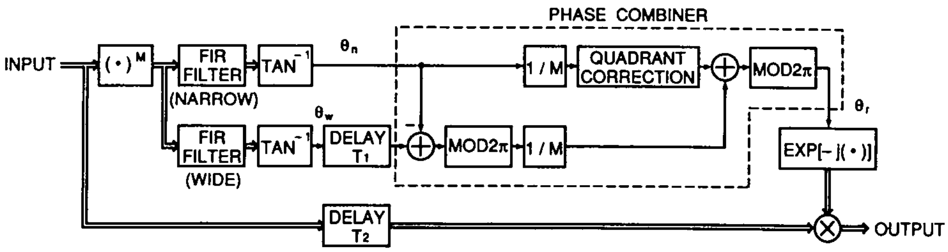

Fig. 2 Proposed carrier recovery block diagram

T_2 in Fig. 2 are determined using baud interval T_b as follows,

$$T_1 = \frac{N_1 - N_2}{2} T_b, \quad T_2 = \frac{N_1 - 1}{2} T_b. \qquad (4)$$

The BER performance for several detection schemes including the proposed scheme is evaluated by computer simulation over the Rician fading channel with carrier-to-multipath ratio (Rician factor) C/M of 10 dB and fading pitch f_d of 1/16 modulation rate for QPSK. These results are shown in Fig. 3.

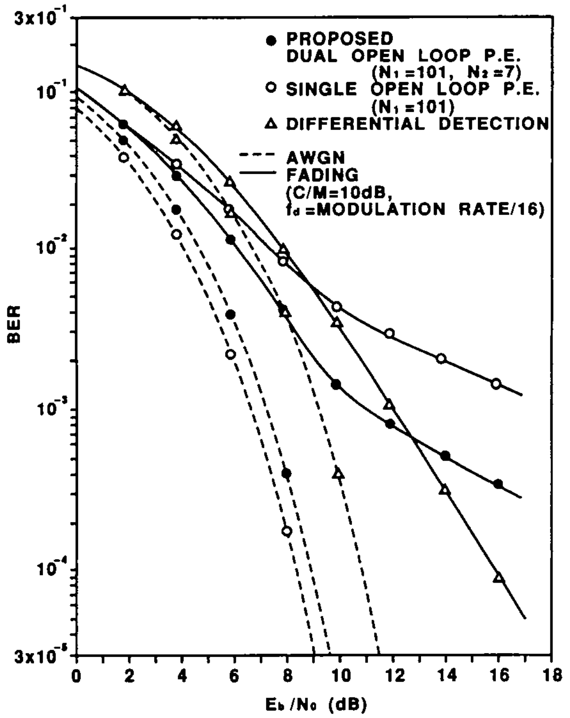

Fig. 3 BER characteristic by simulation

The proposed scheme shows the E_b/N_0 improvement of around 1-2 dB compared with conventional coherent detection using a single open loop phase estimator with narrow bandwidth, and the E_b/N_0 improvement of around 2 dB over non-coherent differential detection, at the operating 1×10^{-2}-5×10^{-3} BER point derived from the high coding rate FEC.

Although static BER performance of the proposed scheme is degraded by the wide band estimator, its amount is relatively small from the conventional scheme, and in an LMS environment its degradation becomes negligible compared with that by fading.

2.2 Quasi-Open Loop Bit Timing Recovery

In bit synchronization, quick recovery from frequent shadowing is another requested characteristic in an LMS environment. To meet such a requirement, authors propose a new bit timing recovery scheme, which is able to make free from acquisition problems by taking quasi-open loop structure. The block diagram is shown in Fig. 4. After root Nyquist filtering, a signal envelope is detected, and then phase correlation between the signal envelope and a complex sinusoidal local reference is calculated. A ROM table is referred to generate the sinusoidal reference at the input sample timing. T_s as sample interval, phase correlation $\mathrm{S}_c(nT_s)$ is expressed by

$$\mathrm{S}_c(nT_s) = V_e(nT_s)\exp(-j2\pi n/N), \qquad (5)$$

where $V_e(nT_s)$ is a signal envelope and N is a sample number per baud interval. The correlated output is fed to a low-pass filter to smooth out the phase jitter, and $\tan^{-1}$ operation of the filter's output is performed. The result represents a bit timing error against the local reference. In a case when a slight frequency difference exists between a modulation clock and the local reference, the slow bit timing change is observed. To track the bit timing continuously, a wide band first-order PLL is introduced. A $2\pi/N$ radian phase step size is sufficient for controlling a digital VCO, because its request is to specify the nearest sample to real bit timing from the input N samples. Hence, the PLL's loop gain K is represented as

$$K = 1/N. \tag{6}$$

The loop gain K is designed large enough that the PLL's bandwidth is sufficiently wide compared with the open loop phase estimator's bandwidth (low-pass filter's bandwidth). Thus, the whole response is solely determined by the low-pass filter's characteristics, showing this loop virtually open.

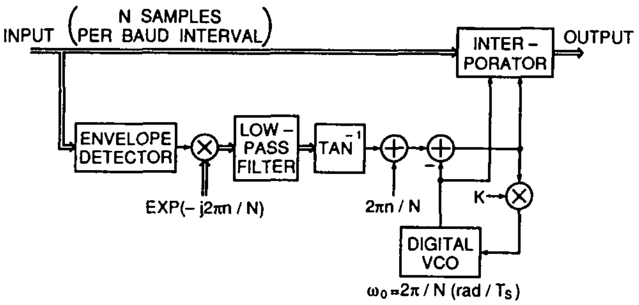

Fig. 4 Bit timing recovery block diagram

A real sampled value at bit timing is obtained by an interporation. The second-order's Lagrange's formula is selected as the interpolation scheme to facilitate the calculation. An interpolated signal $\mathbf{S}_i(r)$ is calculated by using the following equation,

$$\begin{aligned}\mathbf{S}_i(r) &= \frac{r^2-r}{2}\mathbf{S}(-T_s) - (r^2-1)\mathbf{S}(0) + \frac{r^2+r}{2}\mathbf{S}(T_s) \\ &= C_1(r)\mathbf{S}(-T_s) + C_0(r)\mathbf{S}(0) + C_{-1}(r)\mathbf{S}(T_s) \\ &= \sum_{n=-1}^{1} C_{-n}(r)\mathbf{S}(nT_s) \quad (|\, r \,| \leq \frac{1}{2}),\end{aligned} \tag{7}$$

where $\mathbf{S}(0)$ is the nearest sample to real bit timing, $\mathbf{S}(-T_s)$ and $\mathbf{S}(T_s)$ are its adjacent samples, and $r = T_i/T_s$ is normalized interporation timing. This calculation can be implemented by an FIR filter with time variant three tap coefficients $C_i(r)$. These coefficients are written in a ROM table in advance.

3. MODEM IMPLEMENTATION

An experimental terminal modem is developed to evaluate our proposed scheme. From a viewpoint of the spectral efficiency, 5 kHz channel spacing is strongly requested in LMS systems. Therefore, the modulation rate is determined to be 3.2 kbaud for QPSK modulation using band-limit filters with 0.5 roll-off factor. Thus, it enables redundancy for FEC with 3/4 coding rate. Table 1 shows the modem specifications summary.

Items	Specifications
Modulation scheme	3.2 kbaud QPSK
Band - limit filter	Raised cosine roll - off (roll - off factor=0.5)
Operation mode	SCPC continuous mode
Channel spacing	5 kHz
Carrier deviation	Less than ± 3 kHz
A / D sampling rate	606 667 kHz (455 kHzIF x 4 / 3)
A / D quantization	8 bits
Sample number N	6 samples per baud interval
FEC	K=7, R=3 / 4, 6 / 7 Punctured convolutional coding with Viterbi decoding
Voice codec	4.8 kbps multi - pulse coding

Table 1 Modem specifications summary

Figure 5 shows the demodulator block diagram. A 455 kHz intermediate frequency (IF) signal is fed to a single A/D converter, and directly sampled by 4/3×IF clock. Quadrature detection is performed all by digital circuit using these samples, and the digitized orthogonal signal is appropriately decimated. After that, demodulation is performed by three NEC's μPD77230 digital signal processors (DSP) except Viterbi decoding using a dedicated LSI chip, and the following tasks are assigned for each processor.

1) Root Nyquist filtering (FIR filter)
2) Bit timing recovery, AFC and carrier recovery
3) Frame synchronization, de-interleaving, Viterbi decoder interface and Voice decoder interface

Since an open loop phase estimator is very sensitive to a carrier frequency offset, an AFC circuit [3] is necessitated to eliminate this offset before the carrier phase estimation. The AFC circuit is composed of slow and fast tracking parts. The slow tracking AFC part is requested so as to calibrate a local oscillator at all times stably, and the fast tracking AFC part is for the purpose

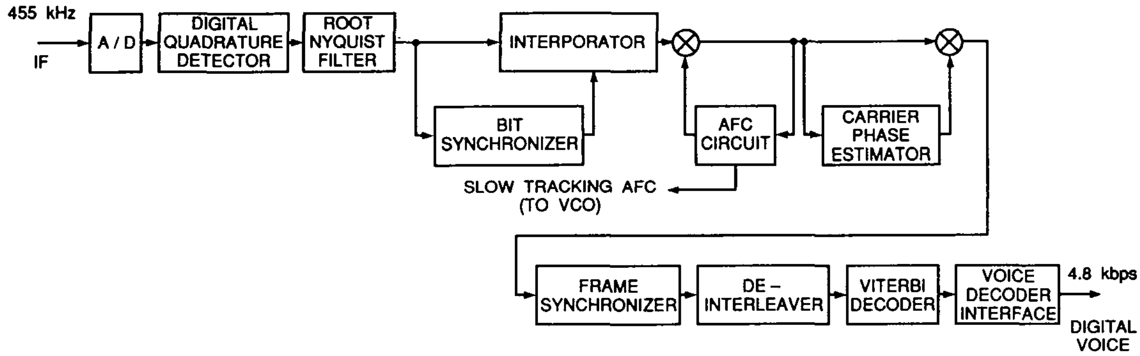

Fig. 5 Demodulator block diagram

of compensating the fast frequency fluctuation by Doppler.

A demodulated signal after Viterbi decoding is sent to a 4.8 kbps voice decoder and converted to a voice signal in it.

4. PERFORMANCE OBTAINED

The modem has been tested using a radio channel simulator. The bit error rate and carrier phase slip rate are measured in the presence of AWGN and Rician fading. The BER performance is shown in Fig. 6 with Rician factor C/M of 10 dB and fading pitch f_d of 200 Hz (1/16 modulation rate). The deviation from computer simulation is well within 1 dB at the operating BER point. To prove the effectiveness of our proposed scheme over the conventional scheme using a single open loop estimator with narrow bandwidth, the performance is measured removing a wide band estimator. As was expected from simulation, the proposed scheme showed the E_b/N_0 improvement of around 1-2 dB over the conventional scheme at the operating 1×10^{-2}-5×10^{-3} BER point. And, a low carrier phase slip rate of less than 5×10^{-2} times/sec is realized at the operating E_b/N_0 of 6 dB with severe Rician fading, as seen in Fig. 7. The demodulated scattering diagrams for the proposed and conventional schemes are depicted in Fig. 8, which shows an appearance that the fast phase fluctuation by fading is effectively compensated by a dual open loop phase estimator.

5. CONCLUSION

A new carrier and bit timing recovery scheme is presented, which remarkably augments the demodulator's performance in an LMS environment. Furthermore, an actual experimental terminal modem has been manufactured to verify the effectiveness of the proposed idea. The results are quite satisfactory, achieving good bit error rate performance and stable synchronization. This scheme is applicable for different modulation schemes.

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REFERENCES

[1] A.J.Viterbi,A.M.Viterbi, "Nonlinear Estimation of PSK-Modulated Carrier Phase with Application to Burst Digital Transmission," IEEE Trans.Inform.Theory, vol.IT-29, pp.543-551, July 1983

[2] W.Hagmann,J.Habermann, "On the phase Error Distribution of an Open Loop Phase Estimator," in Proc.ICC'88, June 1988, pp.1031-1037

[3] F.D.Natali, "AFC Tracking Algorithms," IEEE Trans.Commun., vol.COM-32, pp.935-947, Aug. 1984

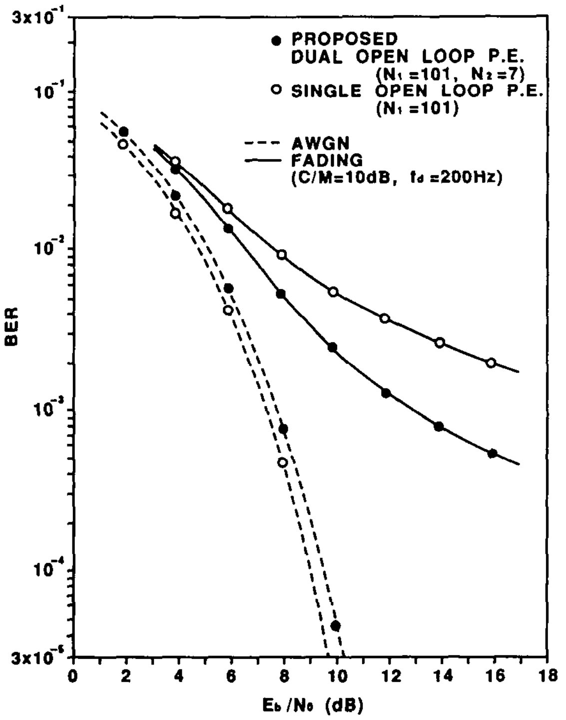

Fig. 6 BER characteristic

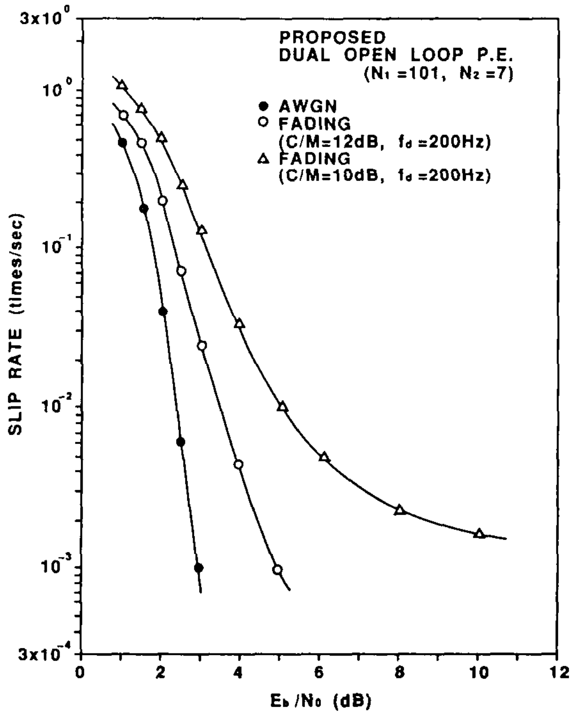

Fig. 7 Carrier slip rate characteristic

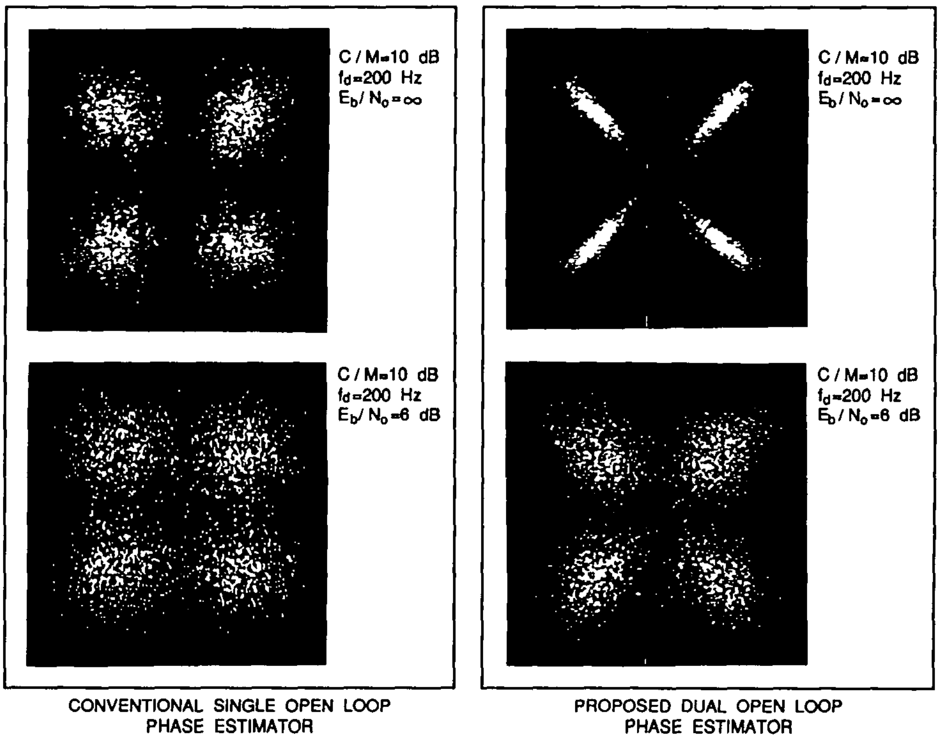

Fig. 8 Demodulated scattering diagrams