Intl J Engg Sci Adv Research 2015 Sep;1(3): 1-6

A novel per-survivor-processing based noncoherent cpm transceiver for ship borne tactical v/uhf waveforms

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ABSTRACT

An FPGA implementation of reduced state Per Survivor Processing (PSP) based GMSK (BT=0.3) transceiver for frequency flat and moderately time varying channels has been envisaged in this paper. The novelty of the work lies in the use of PSP scheme to jointly estimate channel coefficient and phase reference offset for Noncoherent Maximum Likelihood Sequence Detection (MLSE). The communication scenario consists of packet based transmission of Low Data Rate (LDR) 64 kbps waveform in V/UHF band for ship-borne environment. Additionally, frame boundary detection in fading environment is accomplished by using delayed autocorrelation method. Significant improvements have been observed in the receiver performance as compared to our previous work [1] for AWGN environment, where no channel estimation is performed.

Index Terms: Continuous Phase Modulation (CPM), GMSK, Fading, PSP, MLSE, Doppler, Viterbi, FPGA, Synchronization.

1. INTRODUCTION

Invariably for tactical V/UHF band communications, Continuous Phase Modulation (CPM) based modulation schemes are used that fit best in bandwidth and power constrained situations. GMSK (Gaussian Minimum Shift Keying, BT=0.3), one such variant of CPM, offers superior performance in the presence of adjacent channel interference (ACI) and nonlinear amplifiers [2].

In our previous work [1], an FPGA based Noncoherent Sequence Detection scheme has been described for AWGN channels. However, in mobile radio communication systems, the digital mobile radio channels are characterized by time varying multipath propagation which can cause severe performance degradation. Hence, receiver needs to take appropriate measures to mitigate the degradations caused by fading environment.

In our current scenario for ship-borne environment, the speed of aircraft are of the order of 100 nautical miles per hour which offer a maximum Doppler spread of around 100Hz in V/UHF band of 30-400 MHz. Delay spread in this scenario is of the order of 1-2 μsec. Typical data rates of Low Data Rate (LDR) waveforms are of order 64-128 Kbps so this scenario can be treated as frequency flat and moderately time varying channel.

In conventional approach for employing Viterbi Algorithm (VA) under fading conditions, a global channel estimate is used for processing all paths through the trellis [3]. This approach doesn't suits well in dynamically changing channels as it is assumed that channel condition remains constant throughout the packet transmission and hence no channel tracking is employed. In a modified data-aided approach, data estimation is performed on a trellis and tentative decisions from that processing are used for channel estimation. The estimated channel information is then fed back to VA. This approach provides better performance compared to previous, but as it involves feedback of tentative data decisions for channel tracking, the inherent feedback delay causes performance degradation especially in rapidly changing channels[4].

On the contrary, Per Survivor Processing (PSP) updates channel estimate instantaneously for each surviving sequence. This enables estimation and update at the correct time instants across the trellis, thus adapting to rapidly changing channels [3]. Each node updates and maintains its own channel estimate based on its path history. The correct path uses the correct data and the correct history for each step in the trellis estimation process [4]. Even those impairments which are not easy to model and are completely random in nature can be very well tackled using PSP approach. In this paper, to improve the reliability of GMSK receiver in fading environment, a PSP based approach to channel as well as phase reference estimation is employed.

The above approach assumes prior frame synchronization to detect start of packet at the receiver before apply PSP based GMSK demodulation. For packet based transmission, a scheme for frame boundary detection based on delayed autocorrelation method is being employed in the present case. This scheme is capable of achieving frame synchronization in frequency fading environment besides the AWGN only communication scenario.

ISSN NO: 2395-0730

The paper is organized as follows. In Section II, the system model is described along with details of the PSP based sequence detection with channel and noncoherent phase estimation. Packet Structure and delayed autocorrelation based frame synchronization is explained in Section III. Simulation results have been explained in Section IV. FPGA implementation details are given in Section V. Finally concluding remarks are given in Section VI along with scope for future work.

2. SYSTEM MODEL

2.1. Transmitter

The baseband version of GMSK can be expressed as,

$$S(t;\alpha) = \sqrt{\frac{2E}{T}}e^{j\phi(t;\alpha)}$$

where $\phi(t;\alpha) \triangleq 2\pi h \Sigma_{i=-\infty}^{n} q(t-iT)\alpha_i$.

Here α_i is transmitted symbol ,h is modulation index and the phase pulse $q(t)$ is given as,

$$q(t) \triangleq \begin{cases} 0 & t<0 \\ \int_0^t f(\tau)d\tau & 0 \le t < LT \\ 1/2 & t \ge LT \end{cases}$$

which is the integral of the frequency pulse $f(t)$ where $f(t) = \frac{1}{2T}(Q(\gamma B(t-\frac{T}{2})) - Q(\gamma B(t+\frac{T}{2})))$ and $\gamma = 2\pi/\sqrt{log2}$ and $Q(x)$ is the Q-function.

For rational modulation index $h = 2k/p$, the phase can be written as

$$\phi(t;\alpha) = 2\pi h \Sigma_{i=n-L+1}^{n}\alpha_i q(t-iT) \ + \pi h \Sigma_{i=0}^{n-L}\alpha_i$$

where the first term is the correlative phase state (which contains phase contribution due to most recent L symbols) and the second term is phase state (cumulative phase of previous symbols) i.e.

$$\phi(t;\alpha) = \theta(t;\alpha) \ + \theta_{n-L}$$

2.2. Receiver

The received signal under flat and time varying fading environment is expressed as,

$$r(t) = s(t;\alpha)g(t) \ + \eta(t) \tag{1}$$

where $g(t) = a(t)e^{j\mu(t)}$ is the flat fading channel coefficient with Rayleigh distributed magnitude $a(t)$ and uniformly distributed phase $\mu(t)$. The samples $\eta(t)$ are of AWGN with zero mean and variance σ^2.

Each node of VA maintains its own data sequence estimate. The likelihood function for estimated data sequence $\tilde{\alpha}$ which is to be computed separately for each node is given as,

$$\lambda(\tilde{\alpha}) = -\int_{-\infty}^{\infty}[r(t) - s(t;\tilde{\alpha})\tilde{g}(t;\tilde{\alpha})]^2 dt \tag{2}$$

where $\tilde{g}(t;\tilde{\alpha}) = corr(r(t), s(t;\tilde{\alpha}))$ is estimate of channel coefficient per node normalized to unity magnitude.

A weighted approach of channel coefficient estimation is proposed in [5], where channel coefficient is estimated as the weighted sum of current and N previous channel estimates, given as ,

$$\tilde{g}(nT;\tilde{\alpha}) = w_0 corr(r(t), s(t;\tilde{\alpha})) \ + w_1\tilde{g}((n-1)T;\tilde{\alpha}) + \\ \ldots + w_N\tilde{g}((n-N)T;\tilde{\alpha}) \tag{3}$$

where $nT \le t < (n+1)T$ and the weights $\mathbf{w} = [w_0, w_1, \ldots, w_N]$ are given by Yule-walker equation ,

$$\mathbf{w} = \mathbf{R_y^{-1}p}$$

The covariance matrix $\mathbf{R_y}$ and cross correlation vector $\mathbf{p}$ depends on Signal to Noise Ratio (SNR), and the maximum Doppler frequency of channel. These quantities should be known to receiver or they needs to be estimated at receiver, which incurs significantly increased complexity. Hence it has not been attempted in our work and only the current channel estimate has been considered which is termed as standard case in this paper. The simple case of equally weighing the current and previous channel estimates is also considered for comparison purpose.

Maximizing the expression in (2) is equivalent to minimizing,

$$\lambda(\tilde{\alpha}) = Re\int_{-\infty}^{\infty} r(t)s^*(t;\tilde{\alpha})\tilde{g}^*(t;\tilde{\alpha})dt$$

This correlation is computed recursively as,

$$\lambda_{n+1}(E_n) = \lambda_n(S_n) + Re(e^{-j\tilde{\theta}_{n-L}}Z_n(\tilde{\alpha}_n))$$

where the start state S_n can be expressed as an L element vector $S_n = (\tilde{\theta}_{n-L}, \tilde{\alpha}_{n-L+1}, \tilde{\alpha}_{n-L+2}, \ldots, \tilde{\alpha}_{n-1})$ and each branch of trellis can be defined uniquely by the $L+1$ tuple $\delta_n = (\tilde{\theta}_{n-L}, \tilde{\alpha}_{n-L+1}, \tilde{\alpha}_{n-L+2}, \ldots, \tilde{\alpha}_{n-1}, \tilde{\alpha}_n)$. The end state E_n then becomes $E_n = (\tilde{\theta}_{n-L+1}, \tilde{\alpha}_{n-L+2}, \tilde{\alpha}_{n-L+3}, \ldots, \tilde{\alpha}_n)$, λ_n is the cumulative metric for a given state at time index n, $Re(e^{-j\tilde{\theta}_{n-L}})Z_n(\tilde{\alpha}_n))$ is the incremental metric and $Z_n(\tilde{\alpha}_n)$ is the sampled matched filter output over a symbol interval.

$$Z_n(\tilde{\alpha}_n) = \int_{nT}^{(n+1)T} r(t)\tilde{g}^*(t;\tilde{\alpha}_n)e^{-j\tilde{\theta}(t;\tilde{\alpha}_n)}dt \tag{4}$$

The phase state estimate can be computed recursively as,

$$\tilde{\theta}_{n-L} = \tilde{\theta}_{n-L+1} + \pi h\tilde{\alpha}_{n-L}$$

The above model assumes coherent detection with no phase offset at the receiver. In practical scenario, the phase offset is present due to various reasons such as carrier frequency offset, receiver LO phase drift etc. Non coherent detection

[6]assumes that phase offset is present which can be jointly estimated along with channel coefficient in PSP based MLSE receivers. In this scenario the received signal expression (1) can be modified as,

$$r(t) = s(t;\alpha)g(t)e^{j\beta(t)} \ + \eta(t) \tag{5}$$

where $\beta(t)$ the phase offset present in received signal. The modified recursive metric update [7] is given by,

$$\lambda_{n+1}(E_n) = \lambda_n(S_n) + Re(Q_n^*(S_n)e^{-j\tilde{\theta}_{n-L}}Z_n(\tilde{\alpha}_n)) \tag{6}$$

where Q_n is the complex phase reference estimate. This is recursively updated at each symbol time as,

$$Q_{n+1}(E_n) = aQ_n(S_n) \ + (1-a)(e^{-j\tilde{\theta}_{n-L}}Z_n(\tilde{\alpha}_n)) \tag{7}$$

where a is the Forgetting Factor (FF) in the range $0 < a < 1$. The Forgetting Factor signifies the update rate of the older phase estimates.

3. PACKET STRUCTURE AND FRAME BOUNDARY DETECTION

Generally for AWGN environment, frame boundary detection is carried out based on Unique Word (UW) autocorrelation, where a copy of UW reference is correlated with received signal. In AWGN scenario, the maximal length PN sequences are typically used as UW owing to their good autocorrelation properties. However, autocorrelation using stored reference is not suitable in multipath fading environment (current scenario), as UW loses its autocorrelation property due to its multiplication with fading coefficients.

Frame boundary detection in this scenario is accomplished by doing delayed autocorrelation of received sequence with itself.

$$\psi = \sum_{n=0}^{K-1} \frac{x\left[n\right] x\left[n-K\right]^*}{\sqrt{x\left[n\right]^2 \ + x\left[n-K\right]^2}}$$

where $x\left[n\right]$ $n = 0,1,2\ldots K-1$ are K length known sequence repeated two times. For this the frame structure consists of a known sequence used two times. Training sequence should be short enough so that both the sequence experience similar fading and should be long enough so that sufficient correlation gain is obtained. In our implementation, the correlation threshold T_h has been taken as 0.8. The packet structure shown in Fig.1 consists of 10 ms packet with 640 bits. The Known sequence used for frame boundary detection is of 32 bits each and sync pattern which is used for various transmit level control consists of 64 bits.

4. SIMULATION RESULTS

We consider GMSK transceiver with data rate of 64 Kbps for implementation purpose. Traceback depth of PSP based Viterbi Decoder has been taken as 16. Since $q(LT) \sim 0.5$ for pulse length $L = 3$, we can approximate the GMSK (BT=0.3) signal as partial response CPM with a truncated frequency pulse with $L = 3$.

PN Sequence	PN Sequence	Sync Pattern	Payload

Fig. 1. Packet structure employed in 64 Kbps LDR CPM waveform.

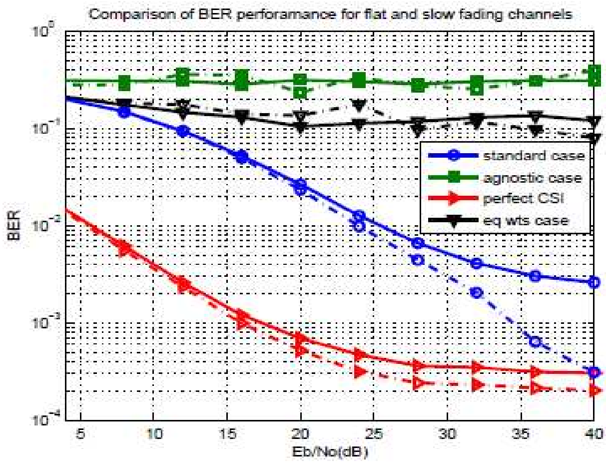

Fig. 2. BER comparison using PSP for flat and slow fading channel. The dotted curves are for FdT=0.001 and solid curves for FdT=0.01.

Firstly we consider coherent detection of transmitted waveform for two cases; slow fading case and moderately time varying channel case. For slow time varying channel scenario, the normalized Doppler offset F_dT has been taken as 0.001 and 0.01. For moderately time varying channel scenario, F_dT has been taken as 0.02 and 0.05. During simulation, we have also considered the case when CSI is known (the ideal situation) as well as the situation when channel impairments are unknown to receiver (the channel uncertainty agnostic case). Also we have considered the case when equal weighted averaging of estimated channel coefficient (as explained in section 2) has been done.

From simulation results in Fig.2 and Fig.3, it can be observed that in scenario when channel is changing dynamically, PSP based channel estimation employing standard averaging (using only current correlation as channel estimate, which accounts to setting $w_0 = 1$ and all the remaining weights zero) provides significant performance improvement over the case when either channel estimation has not been performed at all or the case when channel coefficients are averaged over suc-

ISSN NO: 2395-0730

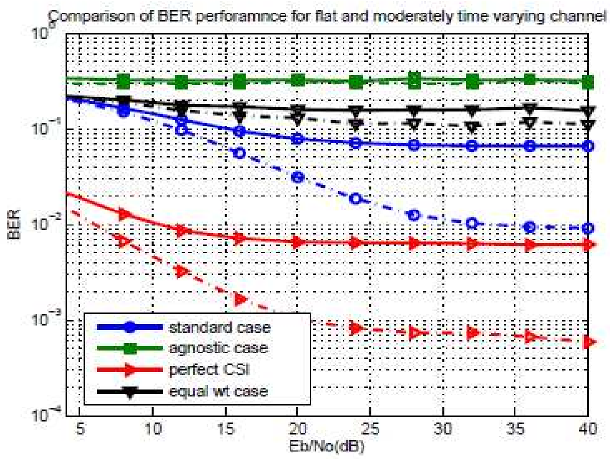

Fig. 3. BER comparison using PSP for flat and moderately time varying channel. Dotted curves are for FdT=0.02 and solid for FdT=0.05

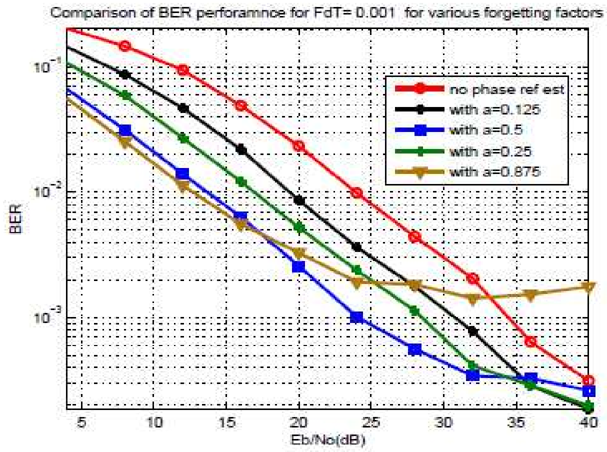

Fig. 4. BER comparison using PSP and employing non coherent phase reference estimate for various of forgetting factor and FdT=0.001.

cessive symbol time epochs with equal weights. It has been reported [5], that when channel is static, equally weighing the channel coefficient over successive time epochs gives slightly better BER performance compared to standard case. It is due to the fact that under static conditions, the channel condition is always the same. Therefore, equally averaging the channel estimates over successive time epochs averages out the noise under static conditions. Using standard approach, PSP based channel estimate under static condition gives slightly inferior performance because of error associated with channel coefficient estimation. This is not the case with time varying channels which is the more general situation we are dealing with. Here standard averaging significantly outperforms equal weight averaging as faster the fading is, faster the channel condition changes, and therefore, putting equal weight on past estimates result in an inaccurate estimation of the channel coefficient.

We then considered the case of Non Coherent Sequence estimation where we are jointly evaluating the channel coefficient as well as phase reference estimates. Simulation results have been presented in Fig.4 and Fig.5 for the normalized Doppler $F_d T$ 0.001 and 0.02 respectively for various values of forgetting factors a. From simulation results, it can be observed that PSP along with joint phase reference and channel estimation gives better BER performance at higher values of a for lower values of E_b/N_0 and at lower values of a for higher E_b/N_0. The forgetting factor $a = 0.5$ comes out to be the optimal choice for wide ranges of E_b/N_0 and hence used in FPGA implementation.

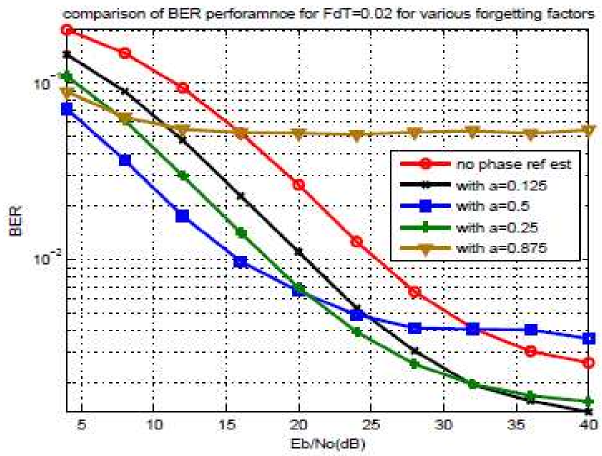

Fig. 5. BER comparison using PSP and employing non coherent phase reference estimate for various of forgetting factor and FdT=0.02.

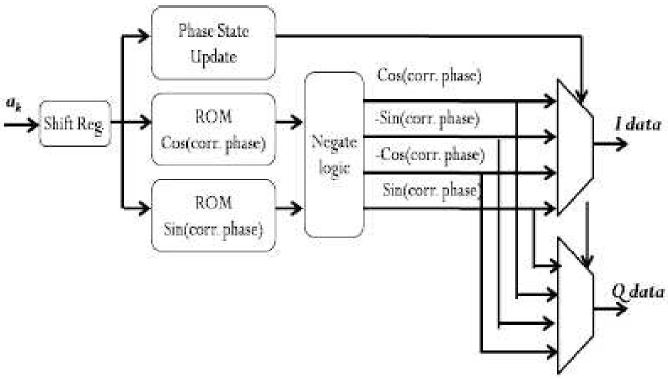

Fig. 6. Baseband GMSK Transmitter

Table 1. I and Q components for different phase states

θ_{n-L}	I_{data}	Q_{data}
0	$cos(\theta(t;\alpha))$	$sin(\theta(t;\alpha))$
$\pi/2$	$-sin(\theta(t;\alpha))$	$cos(\theta(t;\alpha))$
π	$-cos(\theta(t;\alpha))$	$-sin(\theta(t;\alpha))$
$3\pi/2$	$sin(\theta(t;\alpha))$	$-cos(\theta(t;\alpha))$

5. FPGA IMPLEMENTATION

GMSK baseband transceiver is implemented on Xilinx Virtex-5 FPGA using Xilinx System Generator Blocksets [8]. Implementation details of transmitter and receiver are explained in this section.

5.1. Transmitter Architecture

As explained in Section II, phase of GMSK signal is written as,

$$\phi(t;\alpha) = \theta(t;\alpha) + \theta_{n-L}$$

and I and Q data is given as,

$$\begin{aligned} I_{data} &= cos(\theta(t;\alpha) + \theta_{n-L}) \\ Q_{data} &= sin(\theta(t;\alpha) + \theta_{n-L}) \end{aligned} \tag{8}$$

For GMSK(BT=0.3), h=1/2 , M=2 and L=3. Hence, there are $p = 4$ possible phase states $0, \pi/2, \pi, 3\pi/2$. In LUT based implementation, the LUT as given in Table.1 can be formed. Correlative phase states are stored in Read Only Memory (ROM). Fig.6 shows the FPGA model of baseband GMSK transmitter.

5.2. Receiver Architecture

Fig.7 shows Noncoherent PSP based GMSK receiver. Major units of receiver are Metric Calculation and VA units. Metric calculation unit is key part of the PSP based receiver. It performs the functionality of phase reference estimation, channel estimation and incremental metric calculation along with metric update.

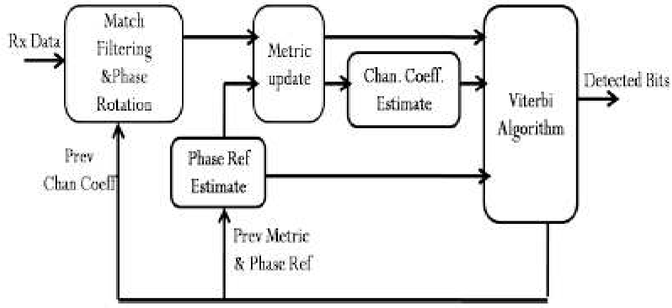

Fig. 7. Baseband GMSK Receiver Block Diagram

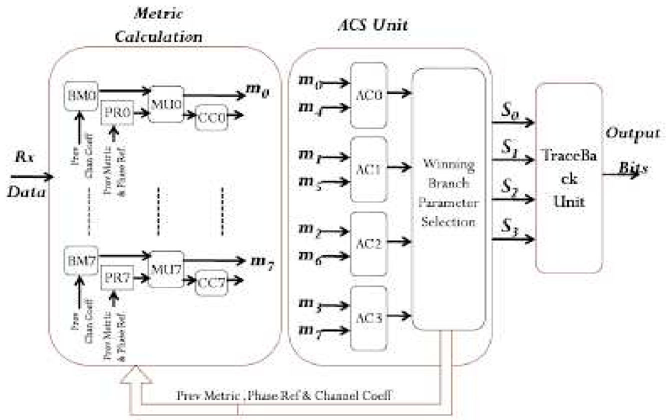

Fig. 8. Baseband GMSK Receiver FPGA Architecture

As shown in Fig.7, current phase reference is estimated using the previous phase reference and previous metric. Current channel coefficient is estimated using current metric and previous channel coefficients. The VA block implements conventional Viterbi algorithm along with feedback logic of winning branch parameters.

Fig.8 shows FPGA implementation architecture of the receiver. Implementation is divided into three major parts. First unit is the metric calculator which comprises of 8 parallel sub blocks. Each sub block computes Final Metric for each node of trellis. BM0 block computes Incremental Metric (4) PR0 block computes current Phase Reference and MU0 block computes final metric (6) by evolving current Phase Reference in the Incremental Metric. CC0 block computes current channel coefficient (3) for each node. Both CC0 block and MU0 block implement a complex multiplier.

The Add-Compare Select (ACS) unit combines the previous node weights and current metrics for the two branches associated with a state, compares the branch metrics and selects the winning branch parameters i.e. channel coefficient , phase reference , metric and phase state.

The Trace Back unit outputs the detected bits. It takes previous states corresponding to the 4 winning states output by the ACS and traces back the states to output the detected bits.

Implementation Architecture of Add-Compare Select (ACS) unit and Trace Back unit is as explained in [1]. The FPGA implementation results have been found to be within 0.2 dB of implementation margin in Xilinx System Generator environment.

6. CONCLUSION

This paper presents a Noncoherent GMSK transceiver design for flat fading and moderately time varying channel environment employing PSP based MLSE approach. Significant performance improvement is obtained by jointly estimating and correcting phase reference estimate along with channel coefficient. An FPGA implementation of transceiver using Xilinx System Generator environment has also been presented.

The scope for further work involves dealing with frequency selective channels and FPGA implementation for the same to cater to the requirements of high data rate waveforms.

Acknowledgement

We thank Director, Defense Electronics Applications Laboratory, Dehradun for permission to publish the paper.

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