A DISTRIBUTED BROADCASTING TIME-SYNCHRONIZATION SCHEME FOR WIRELESS SENSOR NETWORKS

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ABSTRACT

This paper proposes a globally distributed synchronization algorithm for wireless sensor networks. In the proposed scheme, there is a master node that controls the network synchronization and the other sensors periodically broadcast synchronization pulses; they also monitor different frequency channels in an effort to overcome the effect of channel fading.

1. INTRODUCTION

A wireless sensor network is a distributed communications network containing geographically separated sensor nodes. One typical task of sensor networks is to collect local and distributed information and to use low-level sensing to reach global decisions. Thus data fusion plays an important role in comparison with centralized networks, and time synchronization is crucial to enable the fusion of synchronized data. There exist two major approaches to time or clock synchronization [1]. In the centralized approach, there is at least one accurate clock and all other nodes adjust their clocks to it accordingly. In the distributed approach, all clocks achieve synchronization in a distributed manner. Our work starts from the distributed synchronization scheme of [2, 3]. These works assume that the nodes in the network periodically send out synchronization pulses so that each node in the network will see an aggregate pulse. Each synchronization pulse has a zero-crossing within its duration, and the zero-crossing locations are used as synchronization events. The algorithm of [2, 3] assumes a loss-path channel model, but it ignores fading. Due to fading, the signals at a receiving node may combine destructively resulting in a weakened aggregate signal. Moreover, as the number of nodes increases within the coverage area of a receiving node, the received aggregate signal generally tends to zero and the synchronization procedure would then fail. In this paper, we propose a multi-frequency channel approach to overcome the difficulty caused by fading. The synchronization procedure will be based on transmitting reference pulses by each of the sensors and on detecting their times of zero-crossings at the receiving nodes. The effect of channel fading, path loss, and shadowing will be reduced by allowing each sensor to monitor K different frequency channels and by combining the measurements across the channels.

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2. SYSTEM MODEL

We consider a system with N sensor nodes scattered uniformly within a disc of area A. The nodes are labelled from 1 to N. Node 1 is selected as the master clock node, i.e., its clock will serve as the reference clock for the rest of the system. All nodes in the network will seek to synchronize their operational clocks with the master node clock. We assume that all nodes have the same type of clock, but there may exist some random frequency jitters among the clocks. When a new node, or if several new nodes, join the network, they will need to know when they should start their clock pulsing in order to be synchronized with the rest of the nodes in the network.

2.1. Pulse Shaping

The synchronization algorithm will be based on the time of occurrence of certain synchronization events, which should be easy to detect. Since detecting zero-crossings in the analog domain is well studied [4], we may select the zero-crossing within a reference pulse as the synchronization event [2]. An ideal pulse shape would be

$$P(t) = \begin{cases} 1 & -\frac{T}{2} \leq t < 0 \\ 0 & t = 0 \\ -1 & 0 < t \leq \frac{T}{2} \end{cases} \tag{1}$$

where T is the duration of the synchronization pulse. However, the bandwidth required to transmit $p(t)$ is infinite. Alternatively, as discussed in [5] in the context of ultra-wide band communication (UWB), consider the modified Hermite functions

$$P_n(t) = (-1)^n e^{\frac{t^2}{4}} \frac{d^n}{dt^n} \left(e^{-\frac{t^2}{2}} \right) \tag{2}$$

where $n = 0, 1, 2, ...$ and $-\infty < t < \infty$. The first-order modified Hermite function

$$P_1(t) = -te^{-\frac{t^2}{4}} \tag{3}$$

has a desired zero-crossing property. Moreover, as seen in Fig. 3, its pulse shape is smooth in time, has no ringing and the spectrum can be controlled if we modify it as follows:

$$P(t) = \frac{2\sqrt{e}}{T_p} te^{-\left(\frac{\sqrt{2}t}{T_p}\right)^2} \tag{4}$$

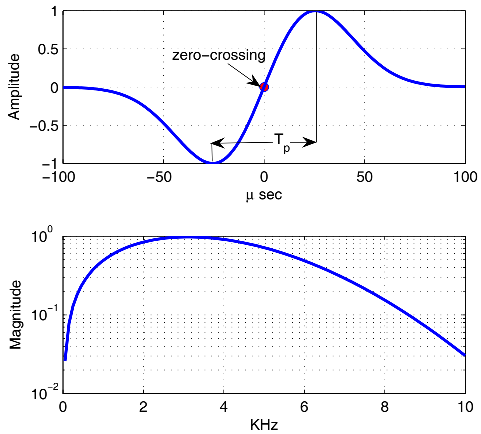

Fig. 1. First-order Hermite function in time (top) and frequency (bottom) used as a reference pulse ($T_p = 51\mu$ sec and $BW = 5$ KHz).

where T_p is the peak-to-peak pulse width. The value of $\sqrt{e}$ is used in order to have unity peak amplitude at $t = T_p$. The corresponding spectrum of this function is given by

$$P(f) = \pi\sqrt{2\pi e}T_p f e^{-\frac{(2\pi f T_p)^2}{2}} \tag{5}$$

which is Gaussian with linear slope in the lower frequencies. The 6 dB bandwidth of the spectrum is $BW \approx 1.6025/2\pi T_p$.

2.2. Channel Model

Two of the factors that influence signal propagation in the wireless medium, and therefore interfere with clock synchronization, are channel fading and path loss shadowing. One of the objectives of the time synchronization scheme proposed in this paper is to reduce the effect of the channel fading condition on the resulting time accuracy. This is achieved by exploiting information about the channel condition in the setup of the algorithm and the synchronization protocol.

- **Channel fading.** When the received signal experiences fading during transmission, both its envelope and phase fluctuate over time. The Rayleigh distribution is frequently used to model channel fading. In this case, the channel gain h_r is distributed as

$$p_{h_r}(r) = \frac{r}{\sigma_h^2} e^{-\frac{r^2}{2\sigma_h^2}} \tag{6}$$

  where σ_h^2 is its variance.

- **Path loss and shadowing.** In addition to the fading effect, the distance between wireless nodes causes path loss in the signal strength. Path loss causes the average received signal power to decrease over distance. Likewise, shadow fading caused by obstacles contributes to the attenuation of the signal strength. The model for path loss and shadowing between two nodes i and j is given by

$$h_s \;=\; \left[PL_0 \left(\frac{d_{i,j}}{d_0}\right)^{\gamma} 10^{\frac{S}{10}}\right]^{-\frac{1}{2}} \tag{7}$$

  where $S \sim N(0, \sigma_s^2)$ and $PL_0 \approx \left(\frac{4d_0}{\lambda}\right)^2$ denotes the path loss at a reference distance d_0. Moreover, γ is the slope of the average increase in the path loss with dB-distance and λ is the wavelength of the transmitted signal. While S is a zero-mean Guassian random variable with standard deviation σ_s and it denotes shadowing and it captures the path loss deviation from its median value.

Based on the above discussion, the received signal at time t by node j in response to a transmission $s(t)$ by node i is given by

$$r(t) = h_r h_s s(t) + v(t) \tag{8}$$

where h_r and h_s are modelled respectively by (6) and (7), and $v(t)$ is additive noise. The average received signal power component by sensor j will be

$$p_r = |h_r|^2 |h_s|^2 p_t \tag{9}$$

where p_t is the transmission power (or variance of $s(t)\times$ energy of pulse shape). We shall assume that p_r stands for the received power over a long period of time, so that the effect of Rayleigh fading can be simplified by replacing $|h_r|^2$ in (9) by $E|h_r|^2$, which is equal to $2\sigma_r^2$, i.e., using (7) and (9) we set

$$p_r \;\approx\; 2\sigma_r^2 \left[PL_0 \left(\frac{d_{i,j}}{d_0}\right)^{\gamma} 10^{\frac{S}{10}}\right]^{-1} p_t \tag{10}$$

3. SYNCHRONIZATION SCHEME

The proposed synchronization procedure will be based on transmitting reference pulses by each of the sensors and on detecting the times of zero-crossings. One way to reduce the effect of channel fading on synchronization accuracy is to allow each sensor in the network to monitor K different frequency bands. Specifically, we make the following assumptions:

1. The propagation delay is negligible due to the close distance of the sensors in comparison with the resolution of the sensors' clocks.
2. The sensors can receive signals in K different frequency channels. A reasonable approximation for the number of required frequency channels is the average number of sensors in the coverage area of a particular sensor, i.e., $N\frac{A_c}{A}$ as shown further ahead after (12).
3. A single frequency channel is assigned to each sensor individually.
4. The transmission frequencies are assigned such that sensors with the same transmission frequency are spread far apart in order to reduce co-channel interference in the coverage area.

4. SENSOR COVERAGE

With the above information, we can estimate the number of frequency bands that are needed within the coverage area A. To do so, we shall first examine the percentage of the area A that will be covered by a single sensor transmission. We shall define the signal coverage for a sensor as the area within which the signal field strength remains above a predefined threshold. To determine the signal coverage, we first evaluate the probability of receiving a signal with signal power above some threshold η at a distance r.

To begin with, equation (10) can be rewritten in dB scale (i.e., using $\bar{x} = 10 \log x$) as

$$\begin{aligned} \bar{p}_r(r) &= \left[\bar{p}_t - \overline{PL}_0 - 10\gamma \log\left(\frac{r}{d_0}\right) + 10\log(2\sigma_h^2) \right] - S \\ &\triangleq p_0(r) - S \end{aligned} \tag{11}$$

where $\bar{p}_r$ and $\bar{p}_t$ are the received and transmitted signal powers in dB scale, respectively. Now since S has a Gaussian distribution, we have

$$P[\bar{p}_r(r) > \eta] = 1 - Q\left(\frac{p_0(r) - \eta}{\sigma_s}\right)$$

where Q is the cumulative distribution function of a zero-mean unit-variance Gaussian random variable. Using (12), we can estimate the coverage area of a sensor as

$$A_c = 2\pi \int_0^R r\left(1 - Q\left(\frac{p_0(r) - \eta}{\sigma_s}\right)\right) dr$$

Additionally, since N sensors are uniformly distributed in the area A, the average number of sensors in the coverage area of each sensor is given by $N\frac{A_c}{A}$.

5. SYNCHRONIZATION ALGORITHM

To describe the synchronization algorithm, we start by noting that the nth transmitted reference signal from the master node at time t will be denoted by

$$s_{1,n}(t) = ap(t - \tau_n) \tag{12}$$

where a is the amplitude of the transmitted signal and τ_n is the time of the nth synchronization event. Moreover, $p(t - \tau_n)$ is the pulse shape from (4) with a zero-crossing at $t = \tau_n$. Each other sensor in the coverage area of the master node will attempt to estimate τ_n and then broadcast a similar pulse to all sensors using its own frequency channel. Specifically, the baseband signal transmitted by an arbitrary sensor i will be

$$s_{i,n}(t) = ap(t - \tau_n - \nu_{i,n}) \tag{13}$$

where $\nu_{i,n}$ is a zero-mean Gaussian jitter with variance σ_ν^2 at the ith node due to errors in estimating the zero-crossing time τ_n and to internal clock jitters. The passband transmitted signal from the ith sensor will be denoted by

$$s_{f_k,i,n}(t) = U_{f_k}\left[s_{i,n}(t)\right] \tag{14}$$

where the operator $U_{f_k}[.]$ stands for upconverting the baseband signal to a passband signal with center frequency f_k. Each sensor will upconvert its signal to a frequency band that is assigned to it individually. The received signal by a generic jth sensor over the kth frequency channel will be

$$r_{f_k,j,n}(t) = h_{i_k,j} s_{f_k,i_k,n}(t) + v_{f_k,j}(t)$$

where i_k is the sensor within the coverage range of sensor j operating in the kth channel, and $v_{f_k,j}(t)$ is the combination of the receiver noise and all interferences at the kth channel from nodes

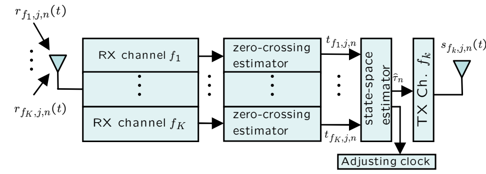

Fig. 2. Synchronization scheme for the jth sensor.

outside the coverage area A_j of node j. Now sensor j will measure K such received signals, one for each of the K frequency channels:

$$\begin{aligned} r_{f_k,j,n}(t) &= h_{i_k,j} ap(t - \tau_n - \nu_{i_k,n}) + v_{f_k,j}(t) \\ & \quad k = 1, ..., K \end{aligned} \tag{15}$$

and proceed to detect the K zero-crossings $t = \tau_n + \nu_{i_k,n}$ – see Fig. 2. Let the noisy estimates of the zero-crossings at the kth channel be denoted by

$$t_{f_k,j,n} = \tau_n + \nu_{i_k,n} + \gamma_{f_k,j,n} \triangleq \tau_n + e_{f_k,j,n} \tag{16}$$

where $\gamma_{f_k,j,n}$ is the detection error due to the receiver noise $v_{f_k,j}(t)$. One simple scheme to estimate τ_n at node j from the K values of $t_{f_k,j,n}$ is to simply average these values, say

$$\hat{\tau}_n = \frac{1}{K}\sum_{k=1}^{K} t_{f_k,j,n} \tag{17}$$

Alternatively, we may set up a state-space model that allows us to exploit better the dynamics of τ_n and its measurements. Thus for each frequency channel k, we shall write

$$\begin{aligned} \tau_{n+1} &= \tau_n + T_c + \bar{T}_c \\ t_{f_k,j,n} &= \tau_n + e_{f_k,j,n} \end{aligned} \tag{18}$$

where T_c is a constant clock step at the master node and $\bar{T}_c$ is a zero-mean Gaussian noise with variance $\sigma_{T_c}^2$ that represents jitter of the master node clock. The K values of all zero-crossing times at node j can be grouped into a single state-space model as

$$\begin{aligned} \tau_{n+1} &= \tau_n + T_c + \bar{T}_c \\ \boldsymbol{t}_{j,n} &= \boldsymbol{a}\tau_n + \Sigma_{j,n}\boldsymbol{e}_{j,n} \end{aligned} \tag{19}$$

where $\boldsymbol{a} \triangleq [1, 1, ..., 1]^T$, $\boldsymbol{e}_{j,n} \triangleq [e_{1,j,n}, ..., e_{K,j,n}]^T$ and $\boldsymbol{t}_{j,n} \triangleq [t_{1,j,n}, ..., t_{K,j,n}]^T$. Moreover, $\Sigma_{j,n}$ is a diagonal weighting matrix with positive entries. Its purpose is to allow us to give more or less weight to zero-crossings that are more or less reliable. One choice for $\Sigma_{j,n}$ would be $\Sigma_{j,n} = \boldsymbol{I}$ in which case all measurements $t_{f_k,j,n}$ are assumed to be subject to similar disturbances around τ_n. Another choice for $\Sigma_{j,n}$ is as follows. Since measurements from the channels with higher power may be asuumed to be more reliable than those from the channels with lower receiving power, we could weight the measurements based on their receiving power. Let p_{r,j,f_k} denote the receiving power at the k channel of node j, i.e.,

$$p_{r,j,f_k} = |h_{i_k,j}|^2 p_t + \sigma_v^2, \quad k = 1, ..., K \tag{20}$$

and introduce the weighting coefficients:

$$a_{j,f_k} = \frac{p_{r,j,f_k} - \sigma_v^2}{\sum_{i=1}^{K}(p_{r,j,f_i} - \sigma_v^2)}, \quad k = 1, ..., K$$

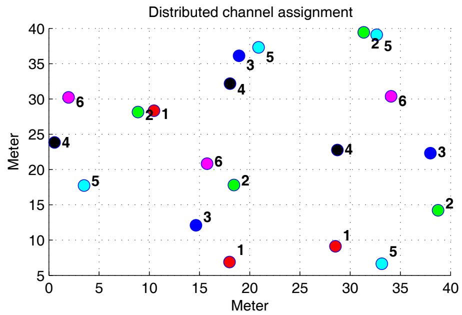

Fig. 3. Distributed channel assignment. Different numbers denote different channels. The channels are distributed uniformly in order to reduce co-channel interference (number of sensors: 20, number of channels: 6, path-loss exponent: 3).

The term $(p_{r,j,f_k} - \sigma_v^2)$ denotes the received signal power, namely, $|h_{i_k,j}|^2 p_t$, which excludes the noise contribution. Then we may set

$$\Sigma_{j,n} = \text{diag}\left\{\frac{1}{\sqrt{a_{j,f_1}}}, ..., \frac{1}{\sqrt{a_{j,f_K}}}\right\} \tag{21}$$

In this way, we are able to give less weight to measurements from channels with large fading or path loss. Now, as is well known, the linear least-mean-squares estimator of τ_n in (19) is given by the Kalman filter [6]:

$$\hat{\tau}_{n+1} = \hat{\tau}_n + \bar{T}_c + \boldsymbol{k}_{p,n}(\boldsymbol{t}_{j,n} - \boldsymbol{a}\hat{\tau}_n) \tag{22}$$

where $\hat{\tau}_0 = 0$, $\boldsymbol{k}_{p,n} = P_n \boldsymbol{a}^* \boldsymbol{R}_{e,n}^{-1}$, $\boldsymbol{R}_{e,n} = \boldsymbol{R} + \boldsymbol{a} P_n \boldsymbol{a}^*$ and the scalar P_n is found recursively via the discrete-time Riccati recursion:

$$P_{n+1} = P_n + \sigma_{T_c}^2 - \boldsymbol{k}_{p,n}\boldsymbol{R}_{e,n}\boldsymbol{k}_{p,n}^*, \ \ P_0 = \pi_0 > 0 \tag{23}$$

Moreover,$\boldsymbol{R} = \Sigma_{j,n} E[\boldsymbol{e}_{j,n}\boldsymbol{e}_{j,n}^*]\Sigma_{j,n}^*$ is the $K \times K$ covariance matrix of the zero-crossing measurement errors. Using (16) we have

$$\boldsymbol{R} = \Sigma_{j,n} E[(\boldsymbol{\nu}_n + \boldsymbol{\gamma}_{j,n})(\boldsymbol{\nu}_n + \boldsymbol{\gamma}_{j,n})^*]\Sigma_{j,n}^* \tag{24}$$

where $\boldsymbol{\nu}_n$ is a $K \times 1$ vector that contains the Gaussian clock jitters from K channels and $\boldsymbol{\gamma}_{j,n}$ is the $K \times 1$ vector that contains the measurement noises from the K channels. We assume the clock jitters from different nodes are independent. We also assume the measurement noises due to interferences and thermal noise are independent from the jitter noises. Then we can write

$$\boldsymbol{R} = (\sigma_\nu^2 + \sigma_\gamma^2).\text{diag}\left\{\frac{1}{a_{j,f_1}}, ..., \frac{1}{a_{j,f_K}}\right\} \tag{25}$$

where σ_ν^2 and σ_γ^2 are the variance of the clocks' jitter noise and measurement noise for each frequency channel and are defined as $\sigma_\nu^2 = E[\nu_n \nu_n^*]$, $\quad \sigma_\gamma^2 = E[\gamma_n \gamma_n^*]$.

6. SYNCHRONIZATION PROTOCOL

Due to frequency reuse and in order to avoid co-channel interference, we propose the following distributed protocol. The channel allocation will be done according to these two steps:

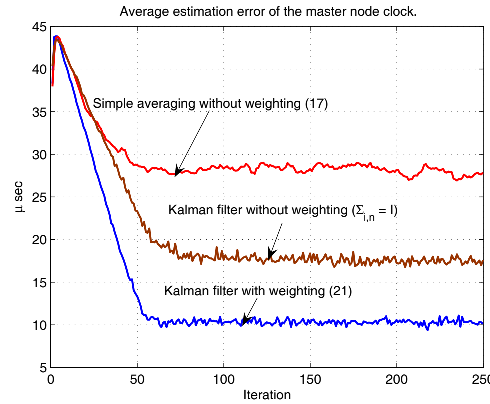

Fig. 4. Average estimation error for 300 nodes ($E|\tau_n - \hat{\tau}_n|$).

- Each node measures the received power in all available channels. Then it chooses the channel with the minimum receiving power as its transmitting channel.
- After selecting its transmitting channel, the node starts transmitting its signal with constant power over that channel.

The first step is a useful condition to reduce interference – see Fig. 3. Each sensor, after receiving reference pulses, starts estimating the times of zero-crossings. Each sensor further estimates the time of the next synchronization event τ_{n+1}, and rebroadcasts the synchronization pulse at this time.

7. SIMULATION

The performance of the proposed scheme has been investigated for a network with 300 nodes and an area of $40 \times 40\, m^2$. The path-loss factor and the shadow fading variance were set to 2.5 and 3.8dB, respectively. We have assumed that the sensors have identical 196 KHz internal clocks and that the synchronization reference pulse is selected to have 1000Hz, 6dB-bandwidth. Moreover, each sensor has 3 channels. Figure 4 shows the average error between the estimated clock time at the sensors and the master node clock for 500 iterations.

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