Dynamic Event-Triggered Fault Detection for Discrete Networked Control System With Time-Delay

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ABSTRACT The dynamic event-triggered fault detection problem for discrete networked control systems with time-delay is investigated in this paper. Firstly, for reducing the triggering times, a new dynamic event-triggered is proposed. Secondly, an observer is constructed on the controller node to generate fault residual, and the mathematical model of the closed-loop systems is established by analyzing the sequence of the transmission signal. Thirdly, by constructing a suitable Lyapunov-Krasovskii functional, the sufficient stability conditions for the closed-loop networked control systems are derived. The computing method of the observer gain matrix, the controller gain matrix and the minimal disturbance repression index is provided. Finally, a case study on the dynamic cart is employed to illustrate the merits of the proposed approach.

INDEX TERMS Time-delay, event-triggered mechanism, networked control system, fault detection.

I. INTRODUCTION Coupled by the progression of communication technology and the increasing scale of control systems, the introduction of computer communication networks into control systems has become a trend in the field of industrial control. The closed-loop feedback systems formed by the shared networks are networked control systems (NCSs) [1], [2], [3], [4]. NCSs own the strength of easily expandable structure, low installation and maintenance costs, etc., and have been widely used in the field of industrial automation [5], [6], [7], [8]. However, some inevitable issues like network-induced time-delay, data packet dropout have appeared due to the introduction of the networks, making system analysis more complicated, and making it more demanding on safety and reliability in engineering applications [9], [10], [11].

With the increasing structure and scale of NCSs, due to the aging of components, external disturbances and changes in the working environment, the NCSs will inevitably have various failures [12], [13]. Therefore, the problem of NCSs fault detection (FD) is a subject of theoretical and practical significance, which has received extensive attention from the academic community and has appeared many results.

The existing literature on FD of NCSs can be divided into three kinds. The first kind literature only considers time-delay. For example, the sum of the data transmission time-delay in the sensor-to-controller (S-to-C) link and the data transmission time-delay in the controller-to-actuator (C-to-A) link was taken as a discrete Markovian chain, thereby NCSs subject to time-delay were modeled by Markovian jump systems (MJSs) [14]. The stability conditions for the closed-loop NCSs were derived. Moreover, the FD observer design method was provided. The S-to-C time-delay and C-to-A time-delay were described with two irreverent discrete Markovian chains respectively, and the NCSs FD method was proposed [15]. The second kind literature only considers data packet dropouts. For instance, assuming that data packet dropout satisfied the Bernoulli distribution character, the FD issue for NCSs subject to S-to-C and C-to-A data packet dropout was researched [16]. The stable conditions for the closed-loop system with a certain robust $H_{\infty}$ disturbance repression performance were obtained and
the FD filter design approach was given. The NCSs with S-to-C and C-to-A data packet dropout were described as MJSs owning four running modes, converting the NCSs FD problem into certain kind of $H_{\infty}$ filter design issue [17]. The third kind literature takes both time-delay and packet dropout into consideration. For example, the conditions for NCSs subject to time-delay and packet dropout owning a certain level of $H_{\infty}$ disturbance repression performance were obtained [18], [19].

Industrial control systems usually sample periodically to control and monitor the controlled plants. However, this traditional periodic sampling technique not only leads to a waste of energy but also increases the communication burden on the network. In this context, the event-triggered mechanisms (ETMs) are proposed, the idea of which is to sample or transmit data only when certain pre-specified conditions are satisfied. Therefore, ETMs can greatly save network bandwidth, thereby help to reduce network-induced time-delay and data packet dropout. Many ETMs have appeared for control or state estimation of various complex systems [20], [21], [22]. According to the different triggering conditions, the ETMs can be mainly divided into static and dynamic ETMs. The static ETMs are currently the most common ones, the triggering threshold of which is a constant [23], [24].

The common triggering condition is as follows:

$$\left[ y_p - y_l \right]^T \left[ y_p - y_l \right] - \mu y_l^T y_l > 0$$

(1)

with $y_p$ being the last transmitted system output, $y_l$ being the current system output and $0 < \mu < 1$ being the triggering threshold. The above condition is based on relative error, and is called relative ETM. The following triggering condition is called absolute event-triggering, because the triggering condition is based on absolute error.

$$\left[ y_p - y_l \right]^T \left[ y_p - y_l \right] - \mu > 0.$$  

(2)

With the purpose of further reducing the triggering times, some dynamic ETMs have been proposed, the characteristic of which is that the triggering threshold can be adjusted according to the evolution of the system, reducing the triggering times and thus saving system resources. Dynamic ETMs can be mainly divided into two types. One is to introduce auxiliary variables described by dynamic equations into the triggering conditions to adjust the triggering threshold [25], [26]. The scalar dynamic equation contains information about the system state, so the obtained threshold changes following the state of the system. The other is that the designed threshold does not evolve according to the dynamic equation, but decays at a certain rate until it reaches a given threshold [27], [28].

From the perspective of FD, the ETMs lose part of the NCSs information due to its non-uniform transmission mode, which increases the difficulty of FD for NCSs. Currently, there are few investigation on the FD of NCSs based on ETMs, and most literatures use the FD filter method. For example, considering the time-delay from the FD object to the FD filter, the problem of FD for NCSs under static ETM based on the FD filter was investigated [29], [30], [31], [32]. By expanding the states of the FD object and the FD filter, the FD issue was converted into a filtering problem, and the design method of the FD filter under ETM was proposed. However, the networks are not only in the S-C channel but also between the C-A channel for typical NCSs. The method proposed by [29], [30], [31], and [32] only took the time-delay or data packet dropout from the FD object to the FD filter into account, and only designed the FD filter, without realizing the co-design of the FD filter and the controller.

To the best of our knowledge, considering both S-to-C and C-to-A time-delay, the co-design of the FD filter and the controller for discrete NCSs under dynamic ETM has not been solved, which is the motivation of this paper. The contributions of this investigation can be exhibited as following points:

1) A new dynamic ETM is proposed, which can effectively reduce the triggering times. Compared with existing methods, the proposed method does not have any restriction on the sampling period.

2) An observer is constructed on the controller node to generate residuals and realize the output feedback control. Through the time sequence analysis of the signal, the closed-loop model of NCSs under the proposed dynamic ETM is established by using the time-delay system approach.

3) A proper Lyapunov-Krasovskii functional which can deal with the S-to-C and C-to-A time-delay independently is constructed. The co-design algorithm of observer gain matrix, the controller gain matrix is provided.

In this paper, through the signal timing analysis under the proposed dynamic ETM, the closed-loop systems subject to S-to-C and C-to-A time-delay are described with equivalent time-delay systems owning two time-delay variables. Furthermore, the co-design approach of the FD observer and the controller is provided by means of dealing with equation inequalities.

The reminder of this paper is arranged as follows. A new dynamic ETM is proposed and the mathematical model of NCSs is established in Section II. Section III provides the co-design approach of the FD observer and the controller. Section IV shows the merits and the effectiveness of the proposed approach by a numerical example. We conclude this paper in Section V.

II. PROBLEM FORMULATION

The considered NCSs’ configuration is illustrated in Figure 1, in which an event generator is installed at the sensor node. The event generator’s function is to decide whether to send system output to the controller at each sampling instant.

The following is the controlled plant’s state space equation:

$$\begin{align*}
    x_{t+1} &= Ax_t + B_d u_t + B_{d\omega} \omega_t + B_{ff} f_t \\
    y_t &= C x_t
\end{align*}$$

(3)
where \( x_l \in \mathbb{R}^n \) represents the plant’s state, \( y_l \in \mathbb{R}^m \) stands for the plant’s output, \( u_l \in \mathbb{R}^m \) means the plant’s control input, \( \omega_l \in \mathbb{R}^k \) indicates the disturbance, \( \tilde{f}_l \in \mathbb{R}^{l_0} \) is the signal of fault. \( A, B_a, B_o, B_f \) and \( C \) are all real matrices with suitable dimensions.

The current sampled output of the system (3) is set as \( y_l \) and the last transmitted sampled output is marked as \( y_{lp}(i = 0, 1, 2, \cdots, \infty) \). The event generator would send the current sampled output \( y_i \), only when \( y_i \) and \( y_{lp}(i = 0, 1, 2, \cdots, \infty) \) satisfy the following proposed dynamic ETM:

\[
[y_{pi} - y_i]^T [y_{pi} - y_i] - \sigma \phi_i - \mu y_i^T y_i > 0,
\]

where \( p_i \in \mathbb{Z}_+ \) is the triggering instant, \( \mu \geq 0 \) is the triggering threshold, \( 0 < \sigma < 0.5 \) is a known scalar and \( \phi_i \) represents an auxiliary variable with the following dynamic equation:

\[
\phi_{i+1} = \sigma \phi_i + \mu y_i^T y_i
\]

with \( \phi_0 \geq 0 \) being the initial condition of \( \phi_i \).

**Remark 1:** Apparently we can see from (4) that, when the \( \sigma \) approaches zero, the dynamic ETM (4) becomes (1), and \( \phi_i \geq 0 \) for \( i \in [0, \infty) \), which means that the triggering times can be decreased effectively compared to that of the static ETM (1).

**Remark 2:** The following dynamic ETM has been proposed in existing literature [33], [34].

\[
[y_{pi} - y_i]^T [y_{pi} - y_i] - \phi_i / \sigma - \mu > 0,
\]

where the internal dynamical variable \( \phi_i \) satisfies

\[
\phi_{i+1} = \delta \phi_i + \mu - [y_{pi} - y_i]^T [y_{pi} - y_i].
\]

with \( \delta \in (0, 1) \) being a given constant.

For \( i \in [p_i + s_{pi} + p_{i+1} + s_{pi+1} - 1] \), one has

\[
[y_i - y_{pi}]^T [y_i - y_{pi}] - \phi_i / \sigma - \mu \leq 0.
\]

It can be obtained that \( \phi_i \geq 0 \) from (6) and (8) on condition that \( \delta > 1 / \sigma \) and \( \phi_0 \geq 0 \). It should be pointed out that \( \phi_i \geq 0 \) only holds for \( i \in [p_i + s_{pi} + p_{i+1} + s_{pi+1} - 1] \), rather than \( i \in [0, \infty) \) since (8) does not hold at the triggering instants.

Let \( s_l \) and \( c_l \) represent S-to-C and C-to-A time-delay respectively, assuming that \( s_l \) and \( c_l \) are both bounded:

\[
s_l \in [0, s_M], c_l \in [0, c_M], \text{ where } s_M \text{ and } c_M \text{ are both non-negative integers.}
\]

Considering the influence of the S-to-C time-delay \( s_l \), the instant when the data released by the event generator \( y_{lp}(i = 0, 1, 2, \cdots, \infty) \) reaches the observer is \( p_i + s_{pi} \). Due to the function of the zero-order holder, the system output received by the observer \( y_i \) can be expressed as:

\[
\tilde{y}_i = y_{pi}, l \in [p_i + s_{pi}, p_{i+1} + s_{pi+1} - 1].
\]

Similar to [35] and [36], consider two cases below:

**Case 1:** If \( p_{i+1} + s_{pi+1} - 1 \leq p_i + 1 + s_M \), define a function \( \eta_i \) as

\[
\eta_i = l - p_i, l \in [p_i + s_{pi}, p_{i+1} + s_{pi+1} - 1].
\]

It can be seen that

\[
s_{pi} \leq \eta_i \leq p_{i+1} - p_i + s_{pi+1} - 1 \leq 1 + s_M.
\]

**Case 2:** If \( p_{i+1} + s_{pi+1} - 1 > p_i + 1 + s_M \), apparently, there exists an integer \( N \geq 0 \) such that

\[
p_i + N + s_M < p_{i+1} + s_{pi+1} - 1 = p_i + 1 + s_M.
\]

It is easy to see that

\[
[p_i + s_{pi}, p_{i+1} + s_{pi+1} - 1] = \{p_i + s_{pi}, p_{i+1} + s_{pi+1} + 1\} \cup \cdots \cup \{p_i + s_{M} + d, p_{i+1} + s_{M} + d + 1\} \cup \{p_i + s_{M} + N, p_{i+1} + s_{M} + N + 1\},
\]

where \( d \in 1, 2, \cdots, N \).

Define \( \eta_i \) as follows

\[
\eta_i = \left\{ \begin{array}{ll} l - p_i, & l \in \Omega_1, \\ l - p_i - d, & l \in \Omega_2, \end{array} \right.
\]

where,

\[
\Omega_1 = [p_i + s_{pi}, p_{i+1} + s_{pi+1} + 1],
\]

\[
\Omega_2 = [p_i + s_{M} + d, p_{i+1} + s_{M} + d + 1], d \in 1, 2, \cdots, N.
\]

From (12), we can obtain

\[
\left\{ \begin{array}{l} \eta_i \leq 1 + s_M \Delta \eta_M, l \in \Omega_1, \\ s_{pi} \leq \Delta \eta_i \leq \eta_M, l \in \Omega_2. \end{array} \right.
\]

For Case 1, define \( h_{i} = 0, l \in [p_i + s_{pi}, p_{i+1} + s_{pi+1} - 1] \).

For Case 2, define

\[
h_{i} = \left\{ \begin{array}{l} 0, l \in \Omega_1, \\ y_{pi} - y_{pi+d}, l \in \Omega_2. \end{array} \right.
\]

From the above definition of \( h_{i} \), the following holds:

\[
y_{pi} = y_i - \eta_i + h_i, \\
l \in [p_i + s_{pi}, p_{i+1} + s_{pi+1} - 1].
\]

Furthermore,

\[
h_i^T h_i \leq \sigma \phi_i + \mu y_i^T y_i - \eta_{i-l}.
\]
At the controller node, construct an observer with the following structure:

\[
\begin{align*}
\dot{x}_{l+1} &= A\hat{x}_l + B_u\bar{u}_l + L (y_{l-\eta} + h_l - \bar{y}_{l-\eta}) \\
\dot{y}_l &= C\hat{x}_l \\
r_l &= W (y_{l-\eta} - \bar{y}_{l-\eta})
\end{align*}
\]  
(17)

where \( \hat{x}_l \in \mathbb{R}^k \) stands for the observer’s state, \( r_l \in \mathbb{R}^l \) means the residual signal and \( \bar{y}_l \in \mathbb{R}^y \) represents the observer’ output. \( L \in \mathbb{R}^{y \times k} \) and \( W \in \mathbb{R}^{q \times y} \) are gain matrices of the observer and the residual, respectively.

Adopt the following observer-based feedback control law:

\[
u_l = K\hat{x}_l.
\]  
(18)

**Remark 3:** Result from the time-delay in C-to-A link, the controlled plant’s control input in (3) differs from the observer control input in (17), and the following holds:

\[
u_l = \tilde{u}_{l-c_l} = K\hat{x}_{l-c_l}.
\]  
(19)

Define the following state estimation error, residual error signal and augmentation vector:

\[
\begin{align*}
e_l &= x_l - \hat{x}_l, e_{rl} &= r_l - \tilde{f}_l, \\
\tilde{e}_l^T &= \left[ e_l \quad 1 \right], \tilde{e}_l^T &= \left[ \omega_l \quad 0 \right],
\end{align*}
\]

Based on the above analysis, when \( l \in [p_l + s_p, p_{l+1} + s_{p_{l+1}} - 1] \), the closed-loop NCS can be expressed by the following time-delay system:

\[
\begin{align*}
\tau_{l+1} &= (A_1 + B_1KI_1)\tau_l + I_2Lh_{l-\eta} + B_2KI_1\tau_{l-c_l} \\
& \quad + B_d\chi_l + I_2LC_1\tau_{l-\eta_l} \\
e_{rl} &= WC_1\tau_{l-\eta_l} - F\chi_l \\
\tau_l &= \varphi_l, l \in \{-\max(m, c_M), \ldots, 0\}
\end{align*}
\]  
(20)

where,

\[
\begin{align*}
A_1 &= \begin{bmatrix} A & 0 \\ 0 & A \end{bmatrix}, B_1 = \begin{bmatrix} 0 \\ -B_u \end{bmatrix}, B_2 = \begin{bmatrix} B_u \\ B_u \end{bmatrix}, \\
B_d &= \begin{bmatrix} B_\omega & B_f \\ B_\omega & B_f \end{bmatrix}, C_1 = \begin{bmatrix} 0 & C \end{bmatrix}, \\
I_1 &= \begin{bmatrix} 1 & -I \end{bmatrix} \in \mathbb{R}^{n \times 2n}, I_2 = \begin{bmatrix} 0 & -I \end{bmatrix}^T \in \mathbb{R}^{2n \times n}, \\
I_3 &= \begin{bmatrix} 0 & I \end{bmatrix} \in \mathbb{R}^{q \times (p+q)}.
\end{align*}
\]

The objective of this paper is to design FD observer (17) and the control law (18) under the proposed dynamic ETM (4) with the consideration of S-to-C and C-to-A time-delay, such that

1) When \( \chi_l = 0 \), the closed-loop NCS (20) is asymptotically stable;

2) The following \( H_\infty \) disturbance suppression performance holds under zero initial conditions.

\[
\sum_{l=0}^{\infty} \tilde{e}_l^T e_{rl} < \gamma^2 \sum_{l=0}^{\infty} \chi_l^T \chi_l
\]  
(21)

with \( \gamma > 0 \) being the disturbance repression performance index. Choose the following evaluation function of residual:

\[
J_l = \sum_{z=\psi_0}^{\psi_{0+l}} \sqrt{r_z^T r_z}
\]  
(22)

The corresponding FD threshold is as follows

\[
J_{th} = \sup_{\chi_l=0} \sum_{z=\psi_0}^{\psi_{0+l}} \sqrt{r_z^T r_z}
\]  
(23)

with \( \psi_0 \) being the initial instant of residual evaluation, and \( L_0 \) being the residual evaluation window length. The FD logic is designed as

\[
\begin{align*}
J_l \leq J_{th} \Rightarrow & \ \text{normal}, \\
J_l > J_{th} \Rightarrow & \ \text{fault}.
\end{align*}
\]

**III. MAIN RESULTS**

We need the following lemma to proceed further.

**Lemma 1** ([37]): For scalars \( \alpha, \alpha_0 \) with \( \alpha \geq \alpha_0 \geq 1 \) and positive definite matrix \( G > 0 \), the following always holds

\[
\sum_{j=\alpha_0}^{\alpha} u_j^T G \sum_{j=\alpha_0}^{\alpha} u_j \leq (\alpha - \alpha_0 + 1) \sum_{j=\alpha_0}^{\alpha} u_j^T G u_j
\]

The following theorem illustrates sufficient conditions for system (20) to be asymptotically stable.

**Theorem 1:** When \( \chi_l = 0 \), for given scalars \( 0 < \mu < 1, 0 < \sigma < 0.5 \), and residual gain matrix \( W \), if there exist positive definite matrices \( P > 0, S_1 > 0, S_2 > 0, Q_1 > 0, Q_2 > 0, Q_1 > 0, Q_4 > 0 \) and matrices \( L, K \) such that

\[
\Gamma = \begin{bmatrix}
\Gamma_{11} & * & * & * & * & * \\
\Gamma_{21} & \Gamma_{22} & * & * & * & * \\
\Gamma_{31} & \Gamma_{32} & \Gamma_{33} & * & * & * \\
\Gamma_{41} & \Gamma_{42} & \Gamma_{43} & \Gamma_{44} & * & * \\
0 & S_2 & 0 & 0 & -Q_4 & -S_2 \\
0 & 0 & S_1 & 0 & 0 & -Q_3 - S_1
\end{bmatrix}
\]  
(24)

where,

\[
\begin{align*}
\Gamma_{11} &= (1 + \phi_M)Q_1 \mu (1 + c_M)Q_2 + (Q_4 - S_1 - S_2 - P + (A_1 + B_1KI_1)^T P (A_1 + B_1KI_1) + \mu (C_4^T C_4), \\
& \quad + \phi_M^2 (A_1 + B_1KI_1)^T S_1 (A_1 + B_1KI_1) - I_1)^T S_1 (A_1 + B_1KI_1 - I_1) + \phi_M^2 (A_1 + B_1KI_1 - I_1)^T S_2 (A_1 + B_1KI_1 - I_1), \\
\Gamma_{21} &= S_2 + (B_2KI_1)^T P (A_1 + B_1KI_1) + \phi_M^2 (B_2KI_1)^T S_1 (A_1 + B_1KI_1) + \phi_M^2 (B_2KI_1)^T S_2 (A_1 + B_1KI_1), \\
\Gamma_{22} &= -Q_2 - 2S_2 + (B_2KI_1)^T P B_2KI_1 \quad + \phi_M^2 (B_2KI_1)^T S_1 B_2KI_1 + \phi_M^2 (B_2KI_1)^T S_2 B_2KI_1, \\
\Gamma_{31} &= S_1 + (I_2LC_1)^T P (A_1 + B_1KI_1) + \phi_M^2 (I_2LC_1)^T S_1 (A_1 + B_1KI_1) + \phi_M^2 (I_2LC_1)^T S_2 (A_1 + B_1KI_1), \\
\Gamma_{32} &= (I_2LC_1)^T P B_2KI_1 + \phi_M^2 (I_2LC_1)^T S_1 B_2KI_1 + \phi_M^2 (I_2LC_1)^T S_2 B_2KI_1,
\end{align*}
\]

and

\[
\begin{align*}
\phi_M &= \frac{1}{2} \max \left( \frac{1}{1 - \mu}, \frac{1}{1 - \sigma} \right). \quad & \phi_L &= \frac{1}{2} \max \left( \frac{1}{1 - \mu}, \frac{1}{1 - \sigma} \right),
\end{align*}
\]

\[
\mu, \sigma > 0,
\]

and

\[
\begin{align*}
\gamma &= \frac{1}{2} \max \left( \frac{1}{1 - \mu}, \frac{1}{1 - \sigma} \right), \quad & \gamma &= \frac{1}{2} \max \left( \frac{1}{1 - \mu}, \frac{1}{1 - \sigma} \right),
\end{align*}
\]

\[
\begin{align*}
\Gamma_{41} &= \Gamma_{42} = \Gamma_{43} = \Gamma_{44} = 0.
\end{align*}
\]
\[ \Gamma_{33} = -Q_1 - 25 + \mu(CI_3) + (I_2LC_1)^T PI_2LC_1 + \eta_3^2(I_2LC_1)^T S_1I_2LC_1 + \sum_{\ell=1} I_2LC_1^T S_1I_2LC_1, \]

\[ \Gamma_{41} = (I_2L)^T P(A_1 + B_1K_1) + \eta_3^2(I_2L)^T S_1(A_1 + B_1K_1) + \sum_{\ell=1} (I_2L)^T S_2A_1 + B_1K_1, \]

\[ \Gamma_{42} = (I_2L)^T PB_2KL_1 + \eta_3^2(I_2L)^T S_1B_2KL_1 + \sum_{\ell=1} (I_2L)^T S_2B_2KL_1, \]

\[ \Gamma_{43} = (I_2L)^T P_2LC_1 + \eta_3^2(I_2L)^T S_1I_2LC_1 + \sum_{\ell=1} (I_2L)^T S_2I_2LC_1, \]

\[ \Gamma_{44} = -I + (I_2L)^T PL_2L + \eta_3^2(I_2L)^T S_1L_2L + \sum_{\ell=1} (I_2L)^T S_2L_2L, \]

\[ I_4 = \left[ 1 \ 0 \right] \in \mathbb{R}^{n \times 2n}, \]

then the NCS (20) is asymptotically stable.

**Proof:** Construct the following Lyapunov-Krasovskii functional:

\[ V_{t_j, \phi_t} = \sum_{j=1}^s V_{t_j} + \phi_t, \quad (25) \]

where

\[ V_{t_j} = t_j^T P_t, \]

\[ V_{t_j} = \sum_{v=1}^{l+1} t_j^T Q_1 t_v + \sum_{v=1}^{l+1} t_j^T Q_2 t_v, \]

\[ V_{t_4} = \sum_{b=\eta_2+1}^{l-1} \sum_{a=\eta_2+1}^{l-1} \eta_2^2 t_j^T Q_1 t_a + \sum_{b=\eta_2+1}^{l-1} \sum_{a=\eta_2+1}^{l-1} \eta_2^2 t_j^T Q_2 t_a, \]

\[ V_{t_5} = \sum_{b=\eta_2+1}^{l-1} \sum_{a=\eta_2+1}^{l-1} \eta_2^2 \sum_{i=1}^{l-1} \sum_{j=i+1}^{l-1} c_{a+b} \epsilon_i \epsilon_j, \]

\[ \epsilon_i = t_i + t_{i+1}. \]

When \( l \in [p_i + s_{p_i} + p_{i+1} + s_{p_{i+1}} - 1], \) along the trajectory of the NCSs (20), we can obtain

\[ \Delta V_{t_i} = t_{i+1}^T P_t t_{i+1} - t_i^T P_t t_i + (A_1 + B_1K_1) t_i + B_2K_1 t_{i-\epsilon_i} + (I_2L)^T P t_i + (A_1 + B_1K_1) t_i + B_2K_1 t_{i-\epsilon_i} - t_i^T P_t t_i \]

\[ = t_i^T ((A_1 + B_1K_1)^T P(A_1 + B_1K_1) - P) t_i + t_i^T (A_1 + B_1K_1)^T PB_2KL_1 t_{i-\epsilon_i}. \]
\[\begin{align*}
\Delta V_{3t_i} &= \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_1 \tau_a + \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_2 \tau_a \\
&\quad - \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_3 \tau_a \\
&\quad - \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_4 \tau_a \\
&= \tau_i^T Q_3 \tau_i - \tau_i^T Q_3 \tau_{i-\eta_{M}} \\
&\quad + \tau_i^T Q_4 \tau_i - \tau_i^T Q_4 \tau_{i-c_{M}}.
\end{align*}\]

(27)

\[\begin{align*}
\Delta V_{4t_i} &= \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_1 \tau_a \\
&\quad + \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_2 \tau_a \\
&\quad - \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_1 \tau_a \\
&\quad + \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \tau_a^T Q_2 \tau_a \\
&= \sum_{b=-\eta_{M}+1}^{l-1} \left( \sum_{a=l+1+b}^{l-1} \tau_a^T Q_1 \tau_a - \sum_{a=l+1+b}^{l-1} \tau_a^T Q_1 \tau_a \right) \\
&\quad + \sum_{b=-c_{M}+1}^{l} \left( \sum_{a=l+1+b}^{l-1} \tau_a^T Q_2 \tau_a - \sum_{a=l+1+b}^{l-1} \tau_a^T Q_2 \tau_a \right) \\
&= \left( \tau_i^T Q_1 \tau_i - \tau_{i+b}^T Q_1 \tau_{i+b} \right) \\
&\quad + \sum_{b=-c_{M}+1}^{l} \left( \tau_i^T Q_2 \tau_i - \tau_{i+b}^T Q_2 \tau_{i+b} \right) \\
&= \eta_{M} \tau_i^T Q_1 \tau_i - \sum_{v=-l-c_{M}+1}^{l} \tau_v^T Q_1 \tau_v \\
&\quad + c_{M} \tau_i^T Q_2 \tau_i - \sum_{v=-l-c_{M}+1}^{l} \tau_v^T Q_2 \tau_v.
\end{align*}\]

(28)

\[\begin{align*}
\Delta V_{5t_i} &= \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \eta_a \epsilon_v^T S_1 \epsilon_a + \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \eta_{a} \epsilon_v^T S_2 \epsilon_a \\
&\quad + \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \eta_a \epsilon_{v}^T S_1 \epsilon_a \\
&\quad - \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \eta_{a} \epsilon_v^T S_2 \epsilon_a \\
&\quad + \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \eta_a \epsilon_{v}^T S_1 \epsilon_a \\
&\quad - \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \eta_{a} \epsilon_v^T S_2 \epsilon_a \\
&= \left( \eta_{M} \epsilon_i^T S_1 \epsilon_i - \eta_{M} \sum_{v=-l-c_{M}}^{l-1} \epsilon_v^T S_1 \epsilon_v \right) \\
&\quad + \left( \sum_{b=-\eta_{M}+1}^{l-1} \sum_{a=l+1+b}^{l-1} \eta_a \epsilon_v^T S_1 \epsilon_a \right) \\
&\quad - \left( \sum_{b=-c_{M}+1}^{l} \sum_{a=l+1+b}^{l-1} \eta_{a} \epsilon_v^T S_2 \epsilon_a \right) \\
&= \eta_{M} \epsilon_i^T S_1 \epsilon_i - \eta_{M} \sum_{v=-l-c_{M}}^{l-1} \epsilon_v^T S_2 \epsilon_v.
\end{align*}\]

(30)

Making use of Lemma 1, we have

\[\begin{align*}
-\eta_{M} \sum_{v=-l-c_{M}}^{l-1} \epsilon_v^T S_1 \epsilon_v - c_{M} \sum_{v=-l-c_{M}}^{l-1} \epsilon_v^T S_2 \epsilon_v \\
&\leq -[\tau_i - \tau_{i-\eta_{M}}]^T S_1 [\tau_i - \tau_{i-\eta_{M}}] \\
&\quad -[\tau_i - \tau_{i-c_{M}}]^T S_2 [\tau_i - \tau_{i-c_{M}}] \\
&\quad -[\tau_{i-\eta_{M}} - \tau_{i-\eta_{M}}]^T S_1 [\tau_{i-\eta_{M}} - \tau_{i-\eta_{M}}] \\
&\quad -[\tau_{i-c_{M}} - \tau_{i-\eta_{M}}]^T S_2 [\tau_{i-c_{M}} - \tau_{i-\eta_{M}}].
\end{align*}\]

(31)

From (26) - (32), under the condition that \(0 < \sigma < 0.5\), we can get

\[\begin{align*}
\Delta V_{t_i, \phi_i + \mu \xi_{i-\eta_{M}}^T, \xi_{i-\eta_{M}}} &\leq \xi_i^T \Gamma \xi_i + (2\sigma - 1)\phi_i \\
&\leq \xi_i^T \Gamma \xi_i,
\end{align*}\]

(33)

where,

\[\begin{align*}
\xi_i &= \left[ \xi_i^T \lambda_i^T \right]^T, \\
\zeta_i &= \left[ \xi_i^T \tau_{i-c_{M}}^T \right]^T.
\end{align*}\]
where
\[
\lambda_I = \left[ h_{II}^T \tau_{I-cm}^T \tau_{I-qm}^T \right]^T.
\]

Therefore, if (24) holds, then the NCS (20) is asymptotically stable, and this completes the proof. □

To obtain the \( K \) and \( L \), the inequality constraints in Theorem 1 should be further processed and we get the following theorem.

**Theorem 2:** When \( \chi_l \neq 0 \), for given scalars \( 0 < \mu < 1 \), \( 0 < \sigma < 0.5 \), and residual gain matrix \( W \), if there exist matrices \( K, L \) and positive definite matrices \( P > 0 \), \( S_1 > 0 \), \( S_2 > 0 \), \( Y > 0 \), \( T_1 > 0 \), \( T_2 > 0 \), \( Q_1 > 0 \), \( Q_2 > 0 \), \( Q_3 > 0 \), \( Q_4 > 0 \) such that

\[
\begin{align*}
\Psi &= \begin{bmatrix} \Psi_{11} & * & * \\ \Psi_{21} & \Psi_{22} & * \\ \Psi_{31} & \Psi_{32} & \Psi_{33} \end{bmatrix} < 0, \\
PY &= I, S_j T_j = I, j \in \{1, 2\},
\end{align*}
\]

where,

\[
\begin{align*}
\Theta_{11} &= \begin{bmatrix} \Theta_{11} & * & * \\ S_2 & -Q_2 - 2S_2 & * \\ S_1 & 0 & -Q_1 - 2S_1 \end{bmatrix}, \\
\Psi_{11} &= \begin{bmatrix} \Theta_{11} & * & * \\ 0 & 0 & 0 \\ 0 & S_2 & 0 \\ 0 & 0 & S_1 \\ 0 & 0 & 0 \end{bmatrix}, \\
\Psi_{22} &= \text{Diag}(-I, -Q_4 - S_2, -Q_3 - S_1, -\gamma^2 I), \\
\Psi_{31} &= \begin{bmatrix} A_1 + B_1 K_l & B_2 K_l & I_2 L C_1 \\ \eta M(A_1 + B_1 K_l) & \eta M B_2 K_l & \eta M I_2 L C_1 \\ c\eta M(A_1 + B_1 K_l) & c\eta M B_2 K_l & c\eta M I_2 L C_1 \\ 0 & 0 & \sqrt{\mu} C I_4 \\ 0 & 0 & WC_1 \\ 0 & 0 & 0 \end{bmatrix}, \\
\Psi_{32} &= \begin{bmatrix} I_2 L & 0 & 0 & B_d \\ \eta M I_2 L & 0 & 0 & \eta M B_d \\ c\eta M I_2 L & 0 & 0 & c\eta M B_d \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}, \\
\Psi_{33} &= \text{Diag}(-Y, -T_1, -T_2, -I_1, -I, -I),
\end{align*}
\]

then the NCS (20) satisfies \( \mathcal{H}_\infty \) performance (21).

**Proof:** when \( \chi_l \neq 0 \), we have

\[
\begin{align*}
\Delta V_{\tau_l, \phi_l} + &\mu \gamma Y_{\tau_l, \phi_l} + \sigma \phi_l - h_{ii}^T h_{il} \\
&\leq \tau_l^T \left( A_1 + B_1 K_l \right)^T P_1 \left( A_1 + B_1 K_l \right) - P \tau_l \\
&+ \tau_l^T \left( A_1 + B_1 K_l \right)^T P B_2 K_l \tau_{l-c_l} \\
&+ \tau_l^T \left( A_1 + B_1 K_l \right)^T P L_2 C_1 \tau_{l-n_l} \\
&+ \tau_l^T \left( A_1 + B_1 K_l \right)^T P L_2 h_{il} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P A_1 \left( A_1 + B_1 K_l \right) \tau_l \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P B_2 K_l \tau_{l-c_l} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P L_2 C_1 \tau_{l-n_l} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P L_2 h_{il} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P B_2 K_l \tau_{l-c_l} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P L_2 C_1 \tau_{l-n_l} \\
&+ \tau_{l-c_l}^T \left( B_2 K_l \right)^T P L_2 h_{il} \tau_{i-c_l}.
\end{align*}
\]
where \( P^{-1} = \begin{bmatrix} 0 & -1 \\ -3 & 0 \end{bmatrix} \). Letting \( P^{-1} = Y, S_1^{-1} = T_1, S_2^{-1} = T_2 \), (34) and (35) can be obtained. Besides, we have

\[
\Delta V_{t_1, \phi_t} + \mu V_{t_1, \phi_t} + \sigma \Phi_1 - K_{11} + \eta_1^T \eta_1 \chi_l < 0.
\]

From (37) and (16), it can be seen that

\[
\Delta V_{t_1, \phi_t} + e_{t_1}^T e_{t_1} - \gamma^2 \chi_l^T \chi_l < 0.
\]

For (38), summing up \( l \) from 0 to \( \infty \), we have

\[
\sum_{l=0}^{\infty} e_{t_l}^T e_{t_l} < \gamma^2 \sum_{l=0}^{\infty} \chi_l^T \chi_l + V_{t_0, \phi_0} - V_{t_\infty, \phi_\infty}.
\]

It can be obtained from the zero initial conditions of the system that

\[
\sum_{l=0}^{\infty} e_{t_l}^T e_{t_l} < \gamma^2 \sum_{l=0}^{\infty} \chi_l^T \chi_l,
\]

which ends the proof. \( \square \)

**Remark 4:** Result from the inverse constraints in (35), we cannot obtain \( L, K \), and \( \gamma_{\min} \) conveniently with the Matlab linear matrix inequality (LMI) toolbox. However, we can convert the inequality constrains in Theorem 2 into the following minimization problem with the aid of the cone complementarity linearization method [38].

\[
\text{Min } Tr \left( PY + \sum_{j=1}^{2} S_j T_j \right) \quad \text{s.t.} (34), (39)
\]

\[
[\begin{bmatrix} P & I \\ I & Y \end{bmatrix}] > 0, \quad [S_j I, T_j] > 0, \quad j \in \{1, 2\}.
\]

The algorithm for computing \( L, K \), and \( \gamma_{\min} \) is given in Algorithm 1.

**IV. A CASE STUDY ON DYNAMIC CART SYSTEM**

In this section, to show the advantage of the proposed dynamic event-triggered FD method, the obtained results will be used to the dynamic cart system [39] shown in Figure 2, where \( m \) means the cart mass, \( u \) stands for the force exerted on the cart, \( s \) represents the cart displacement, \( h \) refers to the buffer viscous friction coefficient, and \( k_s \) is the spring coefficient.

The triggering times under different event-triggered mechanisms are illustrated in Table 1. From Table 1, we can see that the proposed dynamic ETM can efficiently cut the triggering times down, thereby saving the use of computing and communication resources. When \( m = 2 \text{kg}, h = 3 \text{N} \cdot \text{s}/\text{cm}, k_s = 100 \text{N}/\text{cm} \), by selecting the state of the dynamic cart system as \( x = [s, \dot{s}]^T \) and choosing the sampling period as 0.1s, the parameters of the discrete state space model can be obtained as follows

\[
A = \begin{bmatrix} 0.7717 & 0.0853 \\ -4.2658 & 0.6437 \end{bmatrix}, \quad B_u = \begin{bmatrix} 0.0023 \\ 0.0427 \end{bmatrix}.
\]
Algorithm 1 Computing Procedure of $L$, $K$ and $\gamma_{\min}$

1: Let $\gamma = \gamma_0$, set the maximal iterations times $R_{\text{max}}$
2: Obtain a feasible solution $(S^0_1, T^0_1, S^0_2, T^0_2, P^0, Y^0, K^0, L^0)$ satisfying (34) and (39), and set $l = 0$
3: Solve the optimization issue below:
   $$\min \text{ Tr } \left( \sum_{j=1}^2 S_j T_j + PY \right) \text{ s.t. (34) and (39)}$$
4: Set $S^1_1 = S_1$, $T^1_1 = T_1$, $S^1_2 = S_2$, $T^1_2 = T_2$, $P^1 = P$, $Y^1 = Y$, $K^1 = K$, $L^1 = L$
5: while iterations times $< R_{\text{max}}$ do
6:   if (34), (35) hold then
7:      $\gamma = \gamma - \delta$, $l = l + 1$, go to 3th step
8:   else
9:      $l = l + 1$, go to 3th step
10: end if
11: end while
12: if $\gamma < \gamma_0$ then
13:    $\gamma_{\min} = \gamma + \delta$
14: else
15:    No solution can be obtained the optimization problem within the number of iterations $R_{\text{max}}$
16: end if

$C = \begin{bmatrix} 1 & 0 \end{bmatrix}$.

Assuming $B_{\omega} = \begin{bmatrix} 0.005 \\ 0.065 \end{bmatrix}$, $B_f = \begin{bmatrix} 0.003 \\ 0.006 \end{bmatrix}$, the event-triggered threshold is set as $\mu = 0.5$, the residual gain matrix is given as $W = 1$ and $\sigma = 0.3$. Suppose $s_M = c_M = 1$, and the signal of fault is as follows

$$f_l = \begin{cases} 5, & l = 15, \ldots, 30 \\ 0, & \text{otherwise} \end{cases}$$

According to Theorem 2, the observer gain matrix $L$, controller gain matrix $K$, and minimum disturbance suppression
The problem of FD for NCSs under possible network attacks is investigated with consideration of both the impact of the ETM and time-delay based on observer. With the purpose of reducing the triggering times, a new dynamic ETM is proposed. The NCSs subject to both S-to-C and C-to-A time-delay are converted into equivalent time-delay systems. By the method of Lyapunov-Krasovskii functional, the stability conditions of the closed-loop NCSs are established, and the FD observer and the controller gain matrices computation approach is provided, and the co-design of the FD observer and the controller is realized. The problem of FD for NCSs under possible network attacks subject to time-delay and data packet dropout in both S-to-C and C-to-A links with dynamic ETM will be researched in the future.

V. CONCLUSION

The FD issue for NCSs is investigated with consideration of both the impact of the ETM and time-delay based on observer. With the purpose of reducing the triggering times, a new dynamic ETM is proposed. The NCSs subject to both S-to-C and C-to-A time-delay are converted into equivalent time-delay systems. By the method of Lyapunov-Krasovskii functional, the stability conditions of the closed-loop NCSs are established, and the FD observer and the controller gain matrices computation approach is provided, and the co-design of the FD observer and the controller is realized. The problem of FD for NCSs under possible network attacks subject to time-delay and data packet dropout in both S-to-C and C-to-A links with dynamic ETM will be researched in the future.

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TABLE 1. The triggering times under different mechanisms.

| Mechanism            | Triggering times (normal) | Triggering times (fault) |
|----------------------|----------------------------|--------------------------|
| Static mechanism     | 40                         | 37                       |
| Proposed mechanism   | 29                         | 25                       |
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