PAPER Special Section on Digital Signal Processing

Design of IIR Nyquist Filters with Zero Intersymbol Interference

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SUMMARY This paper presents a new method for designing IIR Nyquist filters with zero intersymbol interference. It is shown that IIR Nyquist filters with zero intersymbol interference have some constraints on frequency response, i.e., both magnitude and phase error in passband are dependent on stopband error. Therefore, the frequency response is required to optimize only in stopband. The proposed procedure is based on the formulation of an eigenvalue problem by using Remez multiple exchange algorithm in stopband. Then, the filter coefficients can be computed by solving the eigenvalue problem, and the optimal solution with equiripple stopband response is easily obtained by applying an iteration procedure. The proposed procedure is more computationally efficient than the conventional methods.

key words: *Nyquist filter, IIR filter, eigenvalue problem*

1. Introduction

Nyquist filters play an important role in designing digital transmission systems and filter banks [1]–[10]. Nyquist filters are used to band-limit data spectrum and minimize intersymbol interference. To obtain an exact zero intersymbol interference, every Nth impulse response coefficients is required to be restricted to zero except for one coefficient. Then, they can also be used as efficient decimators and interpolators. There are two kinds of structures for Nyquist filters, FIR filters and IIR filters. Since FIR filters have exact linear phase response and the filter coefficients correspond to an impulse response directly, design of FIR Nyquist filters have been exhaustively studied in [1]–[6]. However, FIR filters generally require high filter order for meeting stringent magnitude specifications. For IIR Nyquist filters, the conventional design requires both time- and frequency-response optimization [7]–[10]. In [7], Nakayama and Mizukami have proposed a class of new transfer functions for IIR Nyquist filters which have exact zero intersymbol interference. Therefore, only the frequency response of the filters is required to optimize by using the proposed transfer function. In [7] and [8], the multistep optimization method is used to optimize the frequency response of IIR Nyquist filters. However, since the optimization algorithm at each step is based on the iterative Chebyshev approximation that applies linear programming techniques at each iteration, it requires heavy computations and the initial value for filter coefficients.

In this paper, we present a new method for designing IIR Nyquist filters with zero intersymbol interference. First, we show that IIR Nyquist filters with zero intersymbol interference have some constraints where the sum of the frequency responses at some related frequencies keep unity regardless of what the values of the filter coefficients are. Therefore, both magnitude and phase error in passband are decided by stopband error, then the frequency response is required to optimize only in stopband. By applying Remez multiple exchange algorithm in stopband, we can formulate the design problem in the form of an eigenvalue problem [11], [12]. Then, the filter coefficients can be computed by solving the eigenvalue problem, and the optimal solution with equiripple stopband response is easily obtained after applying an iteration procedure. The proposed procedure is more computationally efficient than the conventional methods because it has been reduced to computation of the absolute minimum eigenvalue. Finally, we present some design examples to demonstrate the validity of the proposed procedure.

2. Property of IIR Nyquist Filters

To minimize the intersymbol interference, Nyquist filter $H(z)$ is required to have an exact zero crossing impulse response, i.e.,

$$\begin{cases} h(K) = \dfrac{1}{M} & (k=0) \\ h(K+kM) = 0 & (k=\pm1,\pm2,\cdots) \end{cases}, \tag{1}$$

where K and M are integers. It is known in [7] that the transfer function that satisfies the time-domain conditions of Eq. (1) can be expressed in the form

$$H(z) = \frac{z^{-K}}{M} + \frac{\sum_{\substack{i=0 \\ i \neq K+kM}}^{N_n} a_i z^{-i}}{\sum_{i=0}^{N_d} b_i z^{-iM}}, \tag{2}$$

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where N_n, N_d are integers, and the filter coefficients a_i, b_i are real, $b_0 = 1$. Hence, only the frequency-response optimization is required by using the above transfer function. In frequency domain, $H(z)$ is required to be a lowpass filter with linear phase, whose passband and stopband cutoff frequencies ω_p and ω_s are given by

$$\begin{cases} \omega_p = \dfrac{1-\rho}{M}\pi \\ \omega_s = \dfrac{1+\rho}{M}\pi \end{cases}, \tag{3}$$

where ρ is a rolloff rate.

Before approximating the frequency response, we observe the properties of IIR Nyquist filters with zero intersymbol interference. Assume that $H(z)$ is the K-delay version of $\hat{H}(z)$, i.e.,

$$\hat{H}(z) = z^K H(z) = \frac{1}{M} + \frac{\displaystyle\sum_{\substack{i=0 \\ i \neq K+kM}}^{N_n} a_i z^{K-i}}{\displaystyle\sum_{i=0}^{N_d} b_i z^{-iM}}, \tag{4}$$

then the frequency response of $\hat{H}(z)$ can be obtained by

$$\hat{H}(e^{j\omega}) = \frac{1}{M} + \frac{\displaystyle\sum_{\substack{i=0 \\ i \neq K+kM}}^{N_n} a_i \exp\{j(K-i)\omega\}}{\displaystyle\sum_{i=0}^{N_d} b_i \exp\{-jiM\omega\}}. \tag{5}$$

It is well-known that the exponential function satisfies

$$\begin{cases} \displaystyle\sum_{m=0}^{M-1} \exp\left\{ji\left(\omega + \frac{2m\pi}{M}\right)\right\} = 0 & (i \neq kM) \\ \exp\left\{ji\left(\omega + \frac{2m\pi}{M}\right)\right\} = \exp\{ji\omega\} & (i = kM) \end{cases} \tag{6}$$

Therefore, we get

$$\sum_{m=0}^{M-1} \hat{H}(e^{j(\omega_0 + \frac{2m\pi}{M})}) = 1, \tag{7}$$

where $0 \leqq \omega_0 \leqq \omega_p$. Equation (7) means that the sum of the frequency responses at the frequencies $\omega_0 + \frac{2m\pi}{M}$ $(m = 0, 1, \cdots, M-1)$ keep unity regardless of what the values of the coefficients a_i and b_i are. Since $\hat{H}(e^{j(2\pi-\omega)}) = \hat{H}^*(e^{j\omega})$, Eq. (7) can be rewritten as

$$\hat{H}(e^{j\omega_0}) = 1 - \sum_{k=1}^{L-1} \hat{H}(e^{j\omega_k}) - \sum_{k=L}^{M-1} \hat{H}^*(e^{j\omega_k}), \tag{8}$$

where $L = \lfloor \frac{M+1}{2} \rfloor$, $\omega_k = \frac{2k\pi}{M} + \omega_0$ for $0 \leqq k \leqq L-1$, and $\omega_k = \frac{2(M-k)\pi}{M} - \omega_0$ for $L \leqq k \leqq M-1$. x^* and $\lfloor x \rfloor$ denote the complex conjugate and integer part of x, respectively. It is clear from Eq. (8) that if its stopband response is 0, then the frequency response of $\hat{H}(z)$ will be 1, i.e., $H(z) = z^{-K}$ in passband. Therefore, both magnitude and phase error in passband are decided by stopband error. Let δ_s be the maximum magnitude error in stopband, the maximum magnitude and phase error in pasband are

$$\begin{cases} \delta_p \leqq (M-1)\delta_s \\ \Delta\theta \leqq \sin^{-1}(M-1)\delta_s \end{cases}. \tag{9}$$

In practical designs, δ_p and $\Delta\theta$ are usually much smaller than this upper limits. Since δ_p and $\Delta\theta$ are guaranteed to be relatively small for a small value of δ_s, the filter design can concentrate on shaping stopband response. It can also be explained according to the pole-zero locations. $H(z)$ has I $(= N_n + N_d - \lfloor \frac{K}{M} \rfloor - \lfloor \frac{N_n - K}{M} \rfloor)$ independent zeros which are used to provide the desired stopband response. The independent zeros have to locate on or nearby the unit circle to minimize stopband error. The poles and the remaining zeros off the unit circle are used for satisfying the time-domain conditions of Eq. (1) so that passband response is naturally formed. In the following, we will directly apply Remez multiple exchange algorithm in stopband to design IIR Nyquist filters with zero intersymbol interference.

3. Design of IIR Nyquist Filters

In this section, we describe the complex Chebyshev approximation of IIR Nyquist filters with zero intersymbol interference based on the eigenvalue problem.

3.1 Formulation

As shown in Sect. 2, only the frequency response in stopband is required to optimize in design of IIR Nyquist filters with zero intersymbol interference. Since $H(z)$ has I independent zeros to form stopband response, we can select J $(= \lfloor \frac{I}{2} \rfloor + 1)$ extremal frequencies ω_i in stopband as follows;

$$\omega_s = \omega_1 < \omega_2 < \cdots < \omega_J \leqq \pi, \tag{10}$$

where $\omega_J < \pi$ when I is odd because there exists one zero at $\omega = \pi$, and $\omega_J = \pi$ when I is even. We apply Remez multiple exchange algorithm in stopband, and formulate the condition for $\hat{H}(e^{j\omega})$ as follows;

$$\hat{H}(e^{j\omega_i}) = \delta_s e^{j\theta(\omega_i)}, \tag{11}$$

where δ_s is magnitude error and $\theta(\omega)$ is phase response in stopband. Substituting $\hat{H}(e^{j\omega})$ of Eq. (5) into Eq. (11), we divide Eq. (11) into the real and imaginary parts as

$$\sum_{\substack{m=0\\m\neq K+kM}}^{N_n} a_m \cos(K-m)\omega_i + \frac{1}{M}\sum_{m=0}^{N_d} b_m \cos(mM\omega_i) = \delta_s \sum_{m=0}^{N_d} b_m \cos(\theta(\omega_i) - mM\omega_i) \quad (i = 1, 2, \cdots, J), \tag{12}$$

$$\sum_{\substack{m=0\\m\neq K+kM}}^{N_n} a_m \sin(K-m)\omega_i - \frac{1}{M}\sum_{m=0}^{N_d} b_m \sin(mM\omega_i) = \delta_s \sum_{m=0}^{N_d} b_m \sin(\theta(\omega_i) - mM\omega_i) \quad (i = 1, 2, \cdots, J_1), \tag{13}$$

where $J_1 = J$ for odd I, and $J_1 = J-1$ for even I because Eq. (11) has not the imaginary part at $\omega_J = \pi$. Equations (12) and (13) can be rewritten in the matrix form as

$$\boldsymbol{PA} = \delta_s \boldsymbol{QA}, \tag{14}$$

where

$$\boldsymbol{A} = [a_0, \cdots, a_{K-1}, a_{K+1}, \cdots, a_{N_n}, b_0, \cdots, b_{N_d}]^T, \tag{15}$$

$$\boldsymbol{P} = \begin{bmatrix} \cos K\omega_0 & \cdots & \cos\omega_0 & \cos\omega_0 & \cdots & \cos(K-N_n)\omega_0 & \frac{1}{M} & \cdots & \frac{\cos(N_dM\omega_0)}{M} \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ \cos K\omega_J & \cdots & \cos\omega_J & \cos\omega_J & \cdots & \cos(K-N_n)\omega_J & \frac{1}{M} & \cdots & \frac{\cos(N_dM\omega_J)}{M} \\ \sin K\omega_0 & \cdots & \sin\omega_0 & -\sin\omega_0 & \cdots & \sin(K-N_n)\omega_0 & 0 & \cdots & -\frac{\sin(N_dM\omega_0)}{M} \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ \sin K\omega_{J_1} & \cdots & \sin\omega_{J_1} & -\sin\omega_{J_1} & \cdots & \sin(K-N_n)\omega_{J_1} & 0 & \cdots & -\frac{\sin(N_dM\omega_{J_1})}{M} \end{bmatrix} \tag{16}$$

$$\boldsymbol{Q} = \begin{bmatrix} 0 & \cdots & \cos\theta(\omega_0) & \cdots & \cos(\theta(\omega_0) - N_dM\omega_0) \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ 0 & \cdots & \cos\theta(\omega_J) & \cdots & \cos(\theta(\omega_J) - N_dM\omega_J) \\ 0 & \cdots & \sin\theta(\omega_0) & \cdots & \sin(\theta(\omega_0) - N_dM\omega_0) \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ 0 & \cdots & \sin\theta(\omega_{J_1}) & \cdots & \sin(\theta(\omega_{J_1}) - N_dM\omega_{J_1}) \end{bmatrix} \tag{17}$$

It should be noted that Eq. (14) corresponds to a generalized eigenvalue problem, i.e., δ_s is an eigenvalue and $\boldsymbol{A}$ is a corresponding eigenvector. Therefore, in order to minimize the magnitude error δ_s, we must compute the absolute minimum eigenvalue by solving the eigenvalue problem of Eq. (14). Then, the corresponding eigenvector gives a set of filter coefficients. By appropriately selecting the initial extremal frequencies ω_i and its phase $\theta(\omega_i)$, we apply an iteration procedure to attain the optimal solution with equiripple stopband response. The selection of the initial extremal frequencies ω_i and its phase $\theta(\omega_i)$ will directly influence convergence of the iteration procedure. In the following, we will discuss how to select the initial extremal frequencies ω_i and its phase $\theta(\omega_i)$.

3.2 Selection of Initial Value

In the design algorithm, arbitrarily selecting the initial extremal frequencies ω_i and its phase $\theta(\omega_i)$ cannot guarantee to converge to the optimal solution. Hence, it is very important how to select the initial value. It is known that $H(z)$ has I independent zeros in stopband which must locate on or nearby the unit circle. We assume that all the initial zeros locate on the unit circle, i.e., $z_i = e^{\pm j\bar{\omega}_i}$;

$$\omega_s < \bar{\omega}_1 < \bar{\omega}_2 < \cdots < \bar{\omega}_{J_1} \leqq \pi, \tag{18}$$

where $\bar{\omega}_{J_1} = \pi$ when I is odd, and $\bar{\omega}_{J_1} < \pi$ when I is even. A possible choice of $\bar{\omega}_i$ is to pick these frequencies equally spaced in stopband. Other distributions may also be prefered to decrease number of iterations. From Eq. (5), we have

$$\sum_{\substack{m=0\\m\neq K+kM}}^{N_n} a_m e^{j(K-m)\bar{\omega}_i} + \frac{\sum_{m=0}^{N_d} b_m e^{-jmM\bar{\omega}_i}}{M} = 0. \tag{19}$$

By $b_0 = 1$, Eq. (19) can be rewritten into

$$\sum_{\substack{m=0 \\ m \neq K+kM}}^{N_n} a_m \cos(K-m)\bar{\omega}_i + \frac{\sum_{m=1}^{N_d} b_m \cos(mM\bar{\omega}_i)}{M} = -\frac{1}{M} \quad (i = 1, 2, \cdots, J_1), \tag{20}$$

$$\sum_{\substack{m=0 \\ m \neq K+kM}}^{N_n} a_m \sin(K-m)\bar{\omega}_i - \frac{\sum_{m=1}^{N_d} b_m \sin(mM\bar{\omega}_i)}{M} = 0 \quad (i = 1, 2, \cdots, J-1), \tag{21}$$

which is a set of linear equations. By solving the above linear equations, we can obtain a set of filter coefficients, whose independent zeros locate on the unit circle. Then, we can compute the frequency response of $\hat{H}(z)$ by using the obtained filter coefficients, and search for the peak points in stopband to get the initial extremal frequencies ω_i and its phase $\theta(\omega_i)$. Since we have selected the initial locations of I independent zeros all on the unit circle, there must exist J extremal frequencies including the cutoff frequency in stopband. Therefore, The design algorithm is guaranteed to converge to the optimal solution. The design algorithm is shown as follows.

3.3 Design Algorithm

Procedure {Design algorithm of IIR Nyquist filters}

Begin

1. Read N_n, N_d, M, K and ρ.
2. Select initial locations of independent zeros $\bar{\omega}_i$ $(i = 1, 2, \cdots, J_1)$ equally spaced in stopband.
3. Solve Eqs. (20) and (21) to obtain a set of filter coefficients a_i and b_i.
4. Compute frequency response of $\hat{H}(z)$ by using the obtained filter coefficients, then search peak frequencies in stopband as initial extremal frequencies Ω_i $(i = 1, 2, \cdots, J)$ and compute its phase $\theta(\Omega_i)$.

Repeat

5. Set $\omega_i = \Omega_i$ for $i = 1, 2, \cdots, J$.
6. Compute $\boldsymbol{P}$ and $\boldsymbol{Q}$ by using Eqs. (16) and (17), then find the absolute minimum eigenvalue of Eq. (14) to obtain a set of filter coefficients a_i and b_i.
7. Compute frequency response of $\hat{H}(z)$, then search peak frequencies Ω_i $(i = 1, 2, \cdots, J)$ in stopband and compute its phase $\theta(\Omega_i)$.

Until Satisfy the following condition for prescribed small constants ϵ :

$$\{|\Omega_i - \omega_i| \leqq \epsilon \qquad (\text{for } i = 1, 2, \cdots, J)\}$$

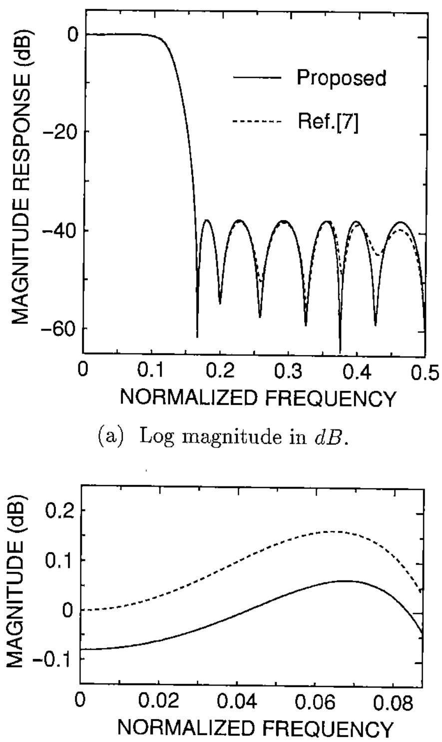
8. Check stability of $H(z)$ by finding the locations of poles.

End.

In the above design algorithm, the obtained filters are not guaranteed to be stable. The stability of $H(z)$ must be checked in step. 8 by computing the locations of the poles. The stability of Nyquist filters are generally dependent on the specifications of the filter, i.e., N_n, N_d, M and K. When numerator order N_n, denominator order N_d and M are given, the group delay must be chosen as $K \geqq K_{\min}$ to guarantee the obtained filters to be stable. $K_{\min}$ is the minimum group delay for the stable filters. In our experience, $K_{\min}$ is directly proportional to N_n and N_d, and inversely to M in general.

4. Design Example

In this section, we present some design examples to demonstrate the validity of the proposed design method,



(a) Log magnitude in dB.

(b) Passband detail.

Fig. 1 Magnitude responses of Example 1.

and compare the performance of the filters with the conventional methods.

Example 1: We consider design of the IIR Nyquist filter of [7] with the following specifications: $N_n = 15$, $N_d = 1$, $K = 9$, $M = 4$ and $\rho = 0.3$ for comparison purposes. The filter is designed by using the proposed procedure. The magnitude response of the resulting filter is shown in Fig. 1, and the group delay is shown in Fig. 2, all in solid line. The passband and stopband attenuations are 0.08 dB and 38 dB, respectively. The maximum error of group delay is 0.65 sampling period in passband. The results in [7] are also shown in dashed line in Fig. 1 and Fig. 2. It is clear that the results of the proposed method and [7] are almost same. The pole-zero locations of two filters obtained from the proposed method and [7] are shown in Fig. 3. It is seen in Fig. 3 that the filter of the proposed method has 13 independent zeros all on the unit circle in stopband, while one of [7] has some zeros nearby the unit circle. To compare computation times, we have executed the proposed algorithm and the algorithm of [7] on SUN SP/IPX. The proposed algorithm required only several seconds while the algorithm of [7] needed more than 10 minutes.

Example 2: We consider design of an IIR Nyquist filter with $N_n = 32$, $N_d = 2$, $K = 19$, $M = 7$ and $\rho = 0.15$. The filter is designed by using the proposed procedure, and the magnitude response and the group delay are shown in Fig. 4 and Fig. 5, respectively. It is seen in Fig. 4 that the magnitude response is equiripple in stopband. We have also designed many filters with various

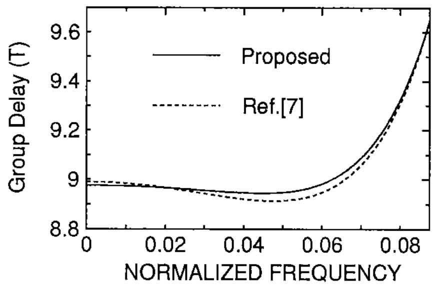
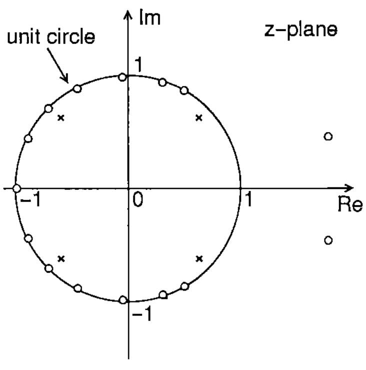
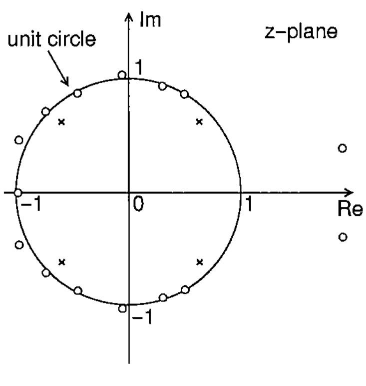


Fig. 2 Group delay of Example 1.

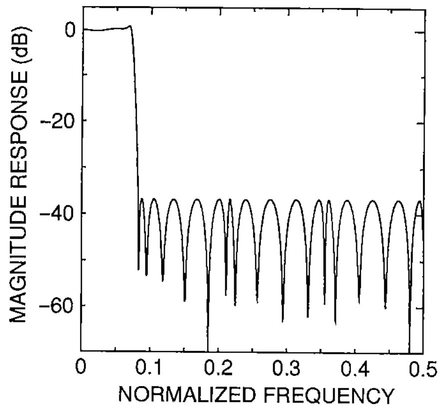


(a) Proposed.

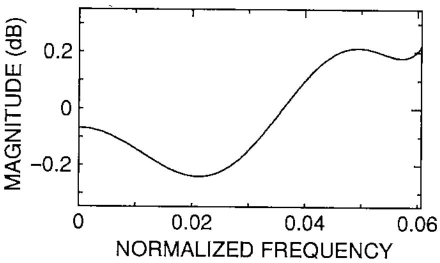


(b) Ref.[7].

Fig. 3 Pole-zero locations of Example 1.



(a) Log magnitude in dB.



(b) Passband detail.

Fig. 4 Magnitude response of Example 2.

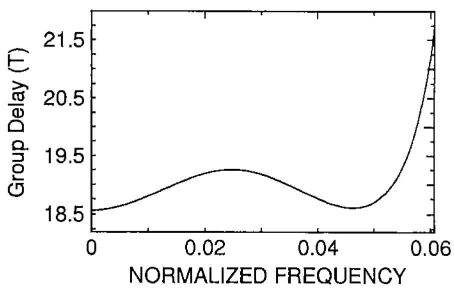


Fig. 5 Group delay of Example 2.

K. We have found that when $K \leqq 18$, the obtained filters have some poles on or outside the unit circle, and will become unstable. Therefore, we have $K_{\min} = 19$, and must set $K \geqq 19$ to obtain the stable filters.

5. Conclusions

In this paper, we have proposed a new method for designing IIR Nyquist filters with zero intersymbol interference. We have shown that IIR Nyquist filters with zero intersymbol interference have some constraints on frequency response where both magnitude and phase error in passband are decided by stopband error. Therefore, the filter design requires to optimize the frequency response only in stopband. The design procedure is based on the formulation of an eigenvalue problem by using Remez multiple exchange algorithm in stopband. Therefore, the filter coefficients can be obtained by computing the absolute minimum eigenvalue, and the optimal solution with equiripple stopband response is easily achieved by applying an iteration procedure. The proposed procedure is more computationally efficient than the conventional methods.

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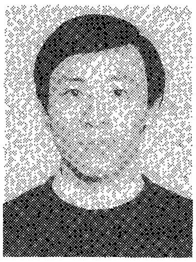

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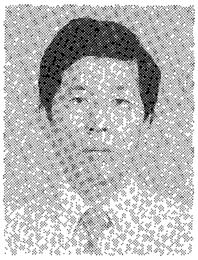

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