Cancellation of Discrete Time Unstable Zeros by Feedforward Control

This paper presents a design methodology for the cancellation of unstable zeros in linear discrete time systems with tracking control objectives. Unstable zeros are defined to be those zeros of the rational transfer function that occur outside the unit circle. Unstable zeros cannot be canceled by feedback without compromising stability. In light of this fact, a feedforward scheme is used. Future desired trajectory information is required because the feedforward scheme is noncausal. The amount of future desired trajectory information that is required depends upon the zero locations and design specifications. It is shown that for a known plant with no zeros on the unit circle one can obtain a frequency response arbitrarily close to 1. Robustness issues and simulation results are discussed.

1 Introduction

The scheme depicted in Fig. 1 is referred to as a two-degree-of-freedom control configuration. An ideal tracking control objective is to have the plant output, \( y(k) \), be equal to the system input \( y_{in}(k) \), i.e., the desired trajectory. For arbitrary inputs this implies the transfer function from input to output be necessarily 1. The feedback controller must be limited to causal filters since the plant is never known precisely, hence the need for feedback in the first place. On the other hand, the feedforward controller can be noncausal if enough future information is known about the desired trajectory. It is precisely this fact that the scheme described below shall exploit. If unstable zeros were uncommon this would not be significant. However, unstable zeros are quite common. Their cause can frequently be attributed to a system delay that is not an integer multiple of the sampling period. In addition, the zero-order-hold equivalent of a continuous time plant with relative degree greater than 2 always yields unstable zeros for sufficiently fast sampling rates (Astrom et al., 1987).

A number of techniques have been developed to minimize the effect of unstable zeros on tracking performance. Among these is the Zero Phase Error Tracking (ZPET) Scheme (Tomizuka, 1987; Shi and Stelson, 1990; Tomizuka, 1989) which is a noncausal feedforward compensation that cancels by plant inversion all poles and specified stable zeros in a stable or previously stabilized system. The essence of the scheme is to also include the inverse of the remaining zeros as zeros in the feedforward compensation block. This has the effect of completely eliminating any phase error. The controller is then adjusted to have unity gain at DC. Note that ZPET can be applied to cancel the phase error caused by zeros on the unit circle and stable zeros that otherwise were not chosen to be canceled by plant inversion. To compensate for the gain error of the plant and that induced by the ZPET controller, Haack and Tomizuka (1991) has developed a systematic means of including additional zeros to reduce the gain error and preserve the zero phase error characteristics. This scheme, referred to as an error or \( E \) filter, is applicable to the case of unstable zeros in the left half plane. For unstable zeros in the right half plane a zero phase lowpass filter can be utilized to reduce excessive feedforward gain at high frequencies. (Menq and Xia, 1990) have characterized unstable phase zeros and developed systematic design procedures for compensating their undesirable gain and phase effects. One can maintain a specified bandwidth with zero phase error and controllable gain errors. Jayasuriya and Tomizuka (1992) has presented a class of feedforward controllers for continuous time systems that in the absence of model uncertainty, noise, and external disturbances, assures perfect tracking for a class of reference signals. Okubo (1987) and Morita et al. (1989) have presented results that utilize the same mechanics as discussed in this paper. In this paper, however, the interpretation is different. The scheme is viewed as an optimal filter in a least squares sense, and frequency weighting is introduced to better tailor the control to the design specifications. Also useful and easy guidelines are derived that make the degree of the approximation to the unstable inverse readily known.

Fig. 1 Feedback/feedforward control system

In this paper, a scheme is described which can make arbitrarily small the tracking degradation effect of unstable zeros. The method may be an alternative to ZPET when a system,
having unstable zeros, requires better gain characteristics. The scheme described here cannot be applied to zeros occurring on the unit circle. The number of required preview steps is dependent on the distance the unstable zero is from the unit circle. The remainder of this paper is organized as follows. Section 2 describes the design procedure. Section 3 presents an optimization point of view. Section 4 addresses robustness. Section 5 presents simulation results and a comparison with ZPET augmented with a zero phase lowpass filter. Section 6 discusses conclusions.

2 Design Procedure

Suppose a known SISO system has been stabilized by some means. By use of a feedforward compensator one can cancel the poles and stable zeros by plant inversion as indicated in Fig. 2. The resulting feedforward compensation can be noncausal but this presents no implementation difficulties provided enough future reference trajectory information is known. The remaining dynamics are those of the uncancellable zeros. We will assume that the uncancellable zeros consist only of unstable zeros. Unstable zeros are defined to be those zeros that occur outside the unit circle. At the end of this section we will consider the case when this assumption is not valid.

Suppose for the sake of illustration a system has only one unstable zero at $-2$ and no zeros on the unit circle. Note, however, that the following is applicable for multiple unstable zeros or zeros that occur as complex conjugate pairs. In Section 5 a real system but with a zero near $+2$ will be described. Suppose the poles and zeros are all cancellable and have been cancelled by feedforward as in Fig. 2. The remaining dynamics are attributable to the zero at $-2$. To cancel the zero at $-2$ we of course cannot include $1/(z+2)$ in the feedforward controller. However, by a series expansion about the origin we have,

$$\frac{1}{z+2} = z^{-1} - 2z^{-1} + 4z^{-2} - 8z^{-3} + 16z^{-4} + \ldots \quad |z| > 2 \quad (1)$$

$$\frac{1}{z+2} = \frac{-1}{4} z^4 + \frac{1}{32} z^8 + \frac{1}{16} z^2 + 1 \quad |z| < 2 \quad (2)$$

Let us suppose that the infinite sequence, Eq. (2), is truncated such that only the first 5 terms are retained. This seems like a reasonable approximation to $1/(z+2)$ since the coefficients of the series are decreasing geometrically. If the 5 term sequence and a one step advance is used as the feedforward filter the input-output transfer function (see Fig. 3) $1 + (z^4/32)$ is obtained. The frequency response can be represented geometrically as in Fig. 4. The frequency response is very close to 1 at all frequencies since it is evaluated for $|z| = 1$.

The location of the zero and the number of terms in the truncated series determine the extent to which a transfer function of 1 is approached. In general, for a real unstable zero at $\alpha$, and with $n$ terms taken from the series, the input-output transfer function will be,

$$1 - \alpha^n \alpha^n$$

It is clear that the gain shall lie within a band

$$1 \pm \frac{1}{\alpha^n} \quad (4)$$

and the maximum value of phase lead and lag, $\theta_{max}$, is given by,

$$\theta_{max} = \arcsin \left( \frac{1}{\alpha^n} \right) \quad (5)$$

So by choosing $n$ such that $\alpha^n$ is large we can get arbitrarily close to 1. Also note that the farther $\alpha$ is from the unit circle the smaller $n$ need be to get a satisfactory frequency response. One may wish to include a factor to correct for the small deviation from unity gain at DC. In this case the factor would be $\alpha^n/(\alpha^n - 1)$. Also note that the gain and phase cycling from max to min occurs $n/2$ times from DC to the Nyquist frequency.

When zeros occur as complex conjugate pairs it is not at first clear how many terms, $n$, one should take. To obtain the coefficients of the feedforward compensator one must invert the complex conjugate pair and then compute the series expansion in positive powers of $z$. To illustrate, suppose our complex pair of unstable zeros are given by the roots of the equation,

$$z^2 + \gamma z + \beta = 0 \quad (6)$$

One must determine the expansion in positive powers of $z$ of,

$$\frac{1}{z^2 + \gamma z + \beta} \quad (7)$$

Note that if we substitute $z^{-1}$ for $z^1$ then the coefficients of the series expansion in positive powers of $z$ are given by the impulse response of the stable filter,

$$\frac{1}{z^{-2} + \gamma z^{-1} + \beta} \quad (8)$$

Resorting to standard Z-transform tables we can readily determine the impulse response which is,
\[ \sum_{n=0}^{\infty} \frac{1}{\beta^{n+1}} \sin((n+1)\omega) \sin(\omega) z^{-n} \] where \( \omega = \arccos \left( \frac{-\gamma}{2B^{\frac{1}{2}}} \right) \)

Since \( \sin((n+1)\omega) \) is bounded by 1 and \( \sin(\omega) \) is a constant, a rough guide as to how many terms should be taken by the factor, \( 1/(\beta^{n+1}) \), i.e., take \( n \) large enough so that this term is negligible. If the unstable phase zeros are expressed as,

\[ (z - \eta)(z - \bar{\eta}) \] (10)

where \( \eta \) and \( \bar{\eta} \) are complex conjugate pairs, the \( n \) term feedforward controller can be directly computed by,

\[ \frac{1}{\beta} \sum_{k=1}^{n} \sum_{i=1}^{n+1-k} \frac{z^{-i+k}}{\eta_{1+i}} \] (11)

Also, in a similar manner to the case of a single real zero, we can compute the feedforward controller as the product,

\[ \left( 1 + \frac{z}{\eta} \right) \ldots \left( 1 + \frac{z}{\eta^{(n)}} \right) \left( 1 - \frac{z}{\eta} \right) \ldots \left( 1 - \frac{z}{\eta^{(n)}} \right) \] (12)

The input output transfer function becomes,

\[ \left( 1 - \frac{z}{\eta} \right) \] (13)

Conservative bounds on the maximum and minimum gain achieved over all frequencies is given by,

\[ 1 + \frac{1}{\eta_{1}} \frac{2}{\eta^{n}} \] (14)

Likewise a conservative bound on phase lead and lag is given by,

\[ \theta_{\text{max}} = 2 \arcsin \left( \frac{1}{\eta_{1}} \right) \] (15)

With this feedforward controller structure the proper choice of \( n \) can be readily ascertained.

If the uncancelable zeros occur on the unit circle then one can define a desired transfer function, \( G_{\text{des}}(z^{-1}) \). The transfer function is constrained in that if the actual plant, \( G_{\text{plan}}(z^{-1}) \), possesses zeros on the unit circle, then \( G_{\text{des}}(z^{-1}) \) must also possess those zeros. This transfer function can be zero phase if so desired, can be noncausal, and will of course embody the desired bandwidth. If zero phase, then \( G_{\text{des}}(z^{-1}) \) has an unstable pole for every stable pole. The unstable poles can be approximated by their series expansion. The feedforward control is then defined to be \( G_{\text{plan}}(z^{-1})^{-1}G_{\text{des}}(z^{-1}) \), where all unstable zeros of \( G_{\text{des}}(z^{-1}) \) are approximated by their series expansion. For mechanical systems large gain demands can result from the cancellation of stable poles and stable zeros. An advantage of this method is the direct specification and achievement of \( G_{\text{des}}(z^{-1}) \).

3 An Optimization Point of View

Again for the sake of simplicity we study the case of a simple zero at \(-2\). The following also applies intoto for zeros that occur as complex conjugate pairs. Consider the following cost index which is a function of the feedforward controller, \( \text{FFD}(z) \).

\[ J(\text{FFD}(z)) = \int_{0}^{2\pi} \left| \frac{1}{z+2} - \text{FFD}(z) \right|^2 d\theta, \quad z = e^{j\theta} \] (16)

From Fourier analysis, if the function \( \text{FFD}(z) \) is restricted to be an \((n-1)\) order polynomial in \( z \), our choice of \( \text{FFD}(z) = 1/(2 - (z/2)) \ldots (z^{-n}/(-2)^{n}) \) is precisely the function that minimizes Eq. (16). This is true since the functions \( z, z^2, \ldots \) are orthogonal with inner product defined as \( \int_{0}^{2\pi} f(z)g(z)dz \) where \( f(z) \) and \( g(z) \) are functions of \( z \) and \( g(z) \) is the complex conjugate of \( f(z) \).

The following can be easily verified,

\[ \int_{0}^{2\pi} \left| \frac{1}{z+2} - \text{FFD}(z) \right|^2 d\theta = \int_{0}^{2\pi} \left| 1 - (z+2)\text{FFD}(z) \right|^2 \left| \frac{1}{z+2} \right|^2 d\theta \] (17)

and

\[ \int_{0}^{2\pi} \left| 1 - (z+2)\text{FFD}(z) \right|^2 d\theta < \kappa \int_{0}^{2\pi} \left| \frac{1}{z+2} - \text{FFD}(z) \right|^2 d\theta \] (18)

where,

\[ \kappa = \sup_{z = e^{j\theta}} |z+2|^2 \] (19)

Note that \( \kappa \) is a positive constant. Therefore, by taking enough terms for the feedforward compensation block the right side of Eq. (18) can be made arbitrarily small and hence the expression on the left side of Eq. (18) can be made arbitrarily small. The expression on the left has clear meaning for the design engineer. It is the integral from zero to twice the Nyquist frequency of the squared magnitude of the difference between the actual frequency response and 1. On closer inspection it is evident that Eq. (17) is the weighted integral of the squared magnitude of the difference between the actual frequency response and 1. As stated before, it is this expression that is in fact minimized by the truncated series expansion. Note that the weighting is greatest at frequencies where the unstable zero has the largest effect on the frequency response. Consequently, the feedforward controller that minimizes this weighted norm may achieve better tracking performance than the controller that minimizes the unweighted norm.

Tracking performance is of course dependent on the designer's chosen means of measuring performance and upon the reference trajectory. If it is known that the reference trajectory has dominant frequency components in a particular band, then frequency weighting can be applied to emphasize this region. For example, suppose we choose a frequency weighting polynomial \( W(z) \). The new cost function we wish to minimize is defined as,

\[ J(\text{FFD}(z)) = \int_{0}^{2\pi} \left| \frac{1}{z+2} - \text{FFD}(z) \right|^2 \left| \frac{W(z)}{z+2} \right|^2 d\theta \] (20)

One can analytically obtain the optimal coefficients of \( \text{FFD}(z) \) in the sense of minimizing Eq. (20). For low order polynomials this is quite easy. For higher order polynomials it lends itself well to a simple computer program. For example, we can emphasize low frequency components if we set \( W(z) = (z + 2)(z + .5) \). The \(-3db\) point occurs at about 120 rad/s. Let \( \text{FFD}(z) \) be chosen to be third order, i.e., \( \text{FFD}(z) = a + bz + cz^2 + dz^3 \). Equation (20) becomes,

\[ J(\text{FFD}(z)) = \int_{0}^{2\pi} \left| 1 - (z+2)(a+bz + cz^2 + dz^3) \right|^2 d\theta \] (21)

The resulting weighted solution is \( \text{FFD}(z) = .4961 - .2399z + .1669z^2 - .0357z^3 \). The input output transfer function with the weighted solution of \( \text{FFD}(z) \) in the feedforward block has a maximum gain variation under 120 rad/s of 3.4 percent and a maximum phase variation of 1.3 degrees. In comparison, the input output transfer function with \( (1/2 - (z/4) + (z^2/8) - (z^3/16) \) in the feedforward block has a maximum gain variation under 120 rad/s of 13 percent and a maximum phase variation of 3.58 degrees.
4 Robustness

A design issue of great importance is that of uncertainty in the location of one or more zeros. We first examine the case of a real unstable zero. Consider that a zero thought to be at \( \eta \) is actually at \( \eta + \delta \). Also assume that \(|\eta + \delta| > 1\), i.e., the actual zero remains unstable despite the uncertainty. The input output transfer function will be,

\[
\left( \frac{-1}{\eta} + \frac{-1}{\eta^2} z + \ldots + \frac{-z^{n-1}}{\eta^{n-1}} \right) (z - (\eta + \delta)) = 1 - \frac{z^n}{\eta^n} + \epsilon(\delta, z), \tag{22}
\]

where \(\epsilon(\delta, z)\), an additive model of the error, represents the difference between the actual transfer function and what is thought to be the transfer function. Solving for \(\epsilon(\delta)\) we obtain,

\[
\epsilon(\delta, z) = \frac{1}{\eta} \left( \frac{1}{\sqrt{\eta}} z + \ldots + \frac{1}{\sqrt{\eta}} z^{n-1} \right) \tag{23}
\]

and for \(z = e^{j\omega}\),

\[
|\epsilon(\delta, z)| \leq \frac{|\delta|}{\eta} \left( \frac{1}{|\sqrt{\eta}|} + \frac{1}{|\sqrt{\eta}|^2} + \ldots + \frac{1}{|\sqrt{\eta}|^{n-1}} \right). \tag{24}
\]

Equation (24) is a tight bound with equality occurring either when \(z = 1\) or \(z = -1\), depending on the sign of \(\eta\). Note that if the zero is on the positive real axis maximum degradation occurs at DC. If the zero is on the negative real axis maximum degradation occurs at the Nyquist frequency. So for a real zero the designer can quickly examine what the maximum gain deviation from 1 is and where it occurs. Noticing that \(1/|\sqrt{\eta}| < 1\) and that \(1/|\sqrt{\eta}| < 1/|\sqrt{\eta}|\) (1/|\sqrt{\eta}|^2), an expedient but conservative bound on the maximum gain deviation is given by,

\[
|\epsilon(\delta, z)| < \frac{|\delta|}{|\sqrt{\eta}|} \left( \frac{1}{|\sqrt{\eta}|} + \frac{1}{|\sqrt{\eta}|^2} + \ldots \right) \leq \frac{|\delta|}{|\eta| - 1}. \tag{25}
\]

Not surprisingly, Eq. (25) indicates that the system tracking performance is much less robust to parameter variations for zeros near the unit circle since \(|\eta| - 1\) can be small. If performance requirements demand good tracking near such a zero and system redesign is not an option, then high gain may also be required (depending on pole locations). High gain must be considered with issues that are not reflected in the transfer function model; however, avoiding high gain in the feedforward block because of the presence of noise is not critical since with the exception of roundoff, there is no source of noise.

For the case of a complex zero the input output transfer function is,

\[
\left( \frac{1}{\eta} \ldots + \frac{1}{\eta^n} z^{n-1} \right) \left( \frac{1}{\eta} \ldots + \frac{1}{\eta^n} z^{n-1} \right) (z - (\eta + \delta))(z - (\eta - \delta)) \tag{26}
\]

which after multiplying yields,

\[
\left( \frac{z^n}{\eta^n} + \frac{\delta}{\eta} \left( \frac{1}{\eta} \ldots + \frac{1}{\eta^n} z^{n-1} \right) \right) \left( \frac{1}{-\eta^n} + \frac{\delta}{\eta} \left( \frac{1}{\eta} \ldots + \frac{1}{\eta^n} z^{n-1} \right) \right) \tag{27}
\]

The additive error term attributable to \(\delta\) is,

\[
\epsilon(\delta, z) = \frac{1}{\eta} \left( \frac{1}{\sqrt{\eta}} z + \ldots + \frac{1}{\sqrt{\eta}} z^{n-1} \right) \left( 1 - \frac{z^n}{\eta^n} \right) + \frac{\delta}{\eta} \left( \frac{1}{\sqrt{\eta}} z + \ldots + \frac{1}{\sqrt{\eta}} z^{n-1} \right) \left( 1 - \frac{z^n}{\eta^n} \right) + \frac{\delta}{\eta} \left( \frac{1}{\sqrt{\eta}} z + \ldots + \frac{1}{\sqrt{\eta}} z^{n-1} \right) \tag{28}
\]

Taking absolute values we obtain the conservative bound,

\[
|\epsilon(\delta, z)| \leq \frac{2|\delta|}{\eta|\eta^n - 1|} + \frac{2|\delta|}{\eta|\eta^n - 1|} + \frac{|\delta|^2}{\eta|\eta^n - 1|^2}. \tag{29}
\]

If we prespecify a bound, \(x\), on \(|\epsilon(\delta, z)|\) we can compute (albeit conservatively) \(rad_{max}\) such that if the actual zero lies in a ball of radius \(rad_{max}\) about \(\eta\) then \(|\epsilon(\delta, z)| < x\). \(rad_{max}\) is given by the positive solution of the quadratic equation,

\[
\frac{2rad_{max}}{\eta|\eta^n - 1|} + \frac{2rad_{max}}{\eta|\eta^n - 1|} + \frac{rad_{max}^2}{\eta|\eta^n - 1|^2} = x. \tag{30}
\]

5 Simulation Results

For comparison purposes simulation results for the truncated series and ZPET with added zero phase lowpass filtering algorithms were obtained for the second stage of a two stage servo system described in (Shi and Stelson, 1990). In Shi and Stelson (1990) the system was modeled as a second order transfer function and the parameters were estimated via a standard recursive least squares algorithm. The transfer function model is,

\[
z(-.215z + .4316) \overline{z^2 + (-1.261)z + .477}} \tag{31}
\]

Note that there are no zeros on the unit circle and an unstable zero is present at 2.007. The zero at 2.007 indicates a reverse reaction process and is most likely due to a structural flexibility.

The ZPET controller for the model is (Tomizuka, 1987),

\[
z^2 + (-1.261)z + .477(\overline{z^2 + 3z^2 + 4.25z^2 + 3z + 1})\overline{.2166z^2} \tag{32}
\]

where one step ahead is needed of the desired trajectory. However, noting that the ZPET algorithm causes large amplification at higher frequencies, we modify the ZPET algorithm by including a zero phase arrangement of high frequency zeros to attenuate the gain. We include zero at \(-.75 \pm .4375j\) and at their respective inverses. Various values were chosen in an ad hoc manner. The overall controller becomes,

\[
z^2 + (-1.261)z + .477(\overline{z^4 + 2z^4 + 4z^4 + 8z^2 + 16z^2 + 32z + 64})\overline{.215(12)z} \tag{33}
\]

Since the poles and stable zero at the origin are also canceled, seven steps ahead are needed of the desired trajectory. The order (numerator) of the ZPET based controller is 7 whereas the order of the truncated series control is 8.

The input output frequency response (magnitude) is shown for both algorithms in Fig. 5. The sampling interval is 10 milliseconds. There are only small differences in performance
\[
\frac{1}{\eta} (1 - \eta^2 z^{-1})
\]

... \[
\text{a conservative bound,}
\]

... \[
\frac{1}{\eta^2} (1 - \eta^2 z^{-1})^2
\]

...\[
\eta < \chi \cdot \text{rad}_\text{max}
\]

...\[
x = \frac{1}{1 - \eta^2}
\]

...\[
x_0
\]

...\[
\text{for the truncated lowpass filtering algorithm of a two stage, 1990). In Shi and second order trans-}
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...\[
\text{standard}
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\text{le and an unstable indicates a reverse structural flexibility, omissions, 2155)
\]

...\[
25z^2 + 3z + 1
\]

...\[
x_0 \approx 2.007
\]

...\[
2 + 32z + 64)
\]

...\[
x_0^2
\]

...\[
\text{for tracking trajectories composed of only low frequency components below about 5 Hz. For precision tracking, however, these small differences may be significant. In addition, if the trajectory is composed of some high frequency components (high speed tracking), the series algorithm can yield significantly improved performance in comparison with the augmented ZPET algorithm. To further demonstrate this point, let the reference trajectory be indicated as in Fig. 6. It is a 1 Hz sawtooth signal of unit amplitude. The control efforts of each algorithm are indicated in Figs. 7 and 8. The series approach generally will make the control effort larger than the ZPET algorithm in the effort to track high frequency components of the reference trajectory. This is not necessarily true for zeros in the right half plane but is true for the much more common occurrence of zeros in the left half plane. Lastly, Figs. 9 and 10 display the tracking error for both schemes. The augmented ZPET filter tracking error is zero everywhere except at a few points per cycle where they are achieving magnitudes of up to \( \pm 0.5 \). The error of the truncated series algorithm is zero for only a few data points per cycle but is consistently small and never actually exceeds bounds of \( \pm 0.015 \).}
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| 1    | In conclusion, the use of ZPET or F-PET for ZPET-based trajectory control or F-PET-based trajectory control is shown to offer better performance. The F-PET-based approach offers an advantage in terms of \[\text{<paste content here>}\]. \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>}\] \[\text{<paste content here>\n
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