Resolver Signal Requirements for High Accuracy Resolver-to-Digital Conversion

DUANE C. HANSELMAN, MEMBER, IEEE

Abstract—Tracking resolver-to-digital (R/D) conversion has emerged as the most robust method for obtaining high-resolution position information from resolvers. When driven by ideal resolver signals, tracking R/D converters currently offer position resolutions up to 2^{16} quantization intervals/period (16-b resolution) and accuracies to 2^{14} intervals/period (14-b accuracy). In this work, the effects of nonideal resolver signal characteristics commonly encountered in practice are investigated. Expressions for the position error reported by an R/D converter due to amplitude imbalance, quadrature error, inductive harmonics, reference phase shift, excitation signal distortion, and disturbance signals are found. From these expressions, bounds on the position accuracy achievable in practical resolver-based position sensing systems are determined.

I. Introduction

Manufacturers of tracking resolver-to-digital (R/D) converters state position resolution and accuracy specifications under the assumption that ideal resolver signals are supplied to the converter. In reality, no resolver generates ideal signals and thus the accuracy specifications of an R/D converter can never be met in practice. In an actual resolver-based position sensing system, amplitude imbalance, quadrature error, inductive harmonics, reference phase shift, excitation signal distortion, and disturbance signals all exist due to the finite precision with which a resolver can be mechanically constructed and electrically excited. In this work, the effects of these nonideal signal characteristics on position accuracy are determined. To isolate the fundamental effects of these signal characteristics from nonideal converter operation itself, the analysis performed assumes ideal R/D conversion. The electronic circuitry and physical operating limitations of tracking R/D converters can be found elsewhere [1]–[3].

Of the nonideal characteristics considered here, only the position error due to reference phase shift and excitation signal distortion have been documented in the past [1], [2]. The position error due to these characteristics have been included here for completeness, and because the derivation of the error expressions is either omitted in past works or is not derived in a way to facilitate meaningful analysis.

This paper proceeds as follows: First, the signals generated by an ideal resolver and the mathematical operation of an ideal tracking R/D converter are reviewed. Second, common nonideal resolver signal characteristics are identified, and expressions for the position error that results from the presence of each nonideal characteristic are derived. Last, the error expressions are analyzed and evaluated to determine the sensitivity of each nonideal characteristic on position accuracy.

Manuscript received January 23, 1990; revised September 5, 1990.
The author is with the Department of Electrical Engineering, University of Maine, Orono, ME 04469.
IEEE Log Number 9040634.

II. Resolvers and R/D Conversion

A. Resolver Construction

A resolver is formed by creating a mechanical structure that has two inductances which are sinusoidal and cosinusoidal functions of position respectively. When an ac current is applied to the structure windings, the voltages induced in these inductances have sinusoidal and cosinusoidal envelopes with respect to position. The position information contained in these two phase voltages are decoded by an R/D converter to give the shaft position of the resolver in digital format.

Resolvers are constructed with either a wound or nonwound rotor. Conventional resolvers are constructed with a wound rotor that is excited with an ac voltage through slip rings and brushes or a rotating transformer [4], [5]. The stator of a conventional resolver has two phases that are in spatial quadrature. As the rotor turns, the mutual coupling between the excited rotor and the stator phases is designed to have the desired sinusoidal and cosinusoidal distributions. A circuit diagram depicting this type of resolver is shown in Fig. 1(a).

Variable-reluctance (VR) resolvers are constructed with a salient, nonwound rotor whose angular position determines the inductance of two stator phases which are in spatial quadrature [6]–[9]. Excitation of the resolver is commonly accomplished by placing additional windings on the stator phases as shown in the circuit diagram in Fig. 1(b). In contrast with the conventional resolver, whose mutual inductance exhibits zero average value, the inductances of a VR resolver are never negative and therefore have a nonzero average value. This fact implies that the voltages induced across the stator windings contain an undesired constant amplitude component in addition to the desired sinusoidal and cosinusoidal envelope waveforms. Before being processed by an R/D converter, this undesired component is eliminated electrically.

0278-0046/90/1200-0556$01.00

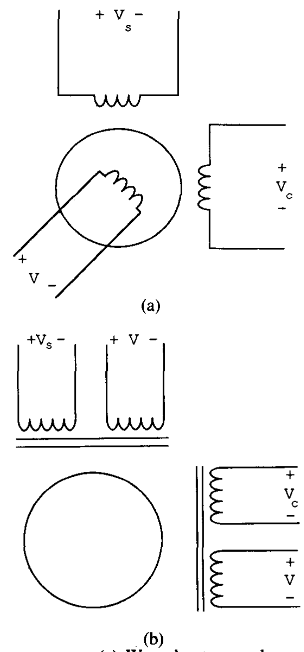

Fig. 1. Resolver structures: (a) Wound rotor resolver, (b) variable-reluctance resolver.

B. Ideal Resolver Operation

With reference to Fig. 1, a resolver is excited by an ac voltage of the form $v = V\cos(\omega t)$. This voltage produces a current in the excitation coils i which in turn induces the output voltages

$$V_s = \frac{d}{dt}\{M_s(\theta)i\}$$

$$V_c = \frac{d}{dt}\{M_c(\theta)i\}, \qquad (1)$$

where, in the ideal case, $M_s(\theta)$ and $M_c(\theta)$ are given by $M_1\sin(\theta)$ and $M_1\cos(\theta)$ respectively, and where θ is the resolver shaft position. No self inductive terms are considered in (1) because the output is assumed to be unloaded and therefore the output current is zero.

In the ideal case when the resistance of the excitation coils is zero, the phase of the output voltage is identical to that of the excitation voltage. Using this information and performing the differentiation indicated in (1) results in the resolver voltages

$$V_s = V_1\sin\theta\cos\omega t + \frac{\dot{\theta}}{\omega}V_1\cos\theta\sin\omega t$$

$$V_c = V_1\cos\theta\cos\omega t - \frac{\dot{\theta}}{\omega}V_1\sin\theta\sin\omega t, \qquad (2)$$

where $V_1 = VM_1/L$ and L is the total inductance of the excitation coils. The first term in these expressions is the desired resolver signal; whereas the second is a speed voltage term. Since this second term is zero when the resolver is at rest, it does not affect static position accuracy. However, under dynamic conditions the affect of the speed voltage term must be considered. In the following, the resolver is assumed to be at rest unless stated otherwise.

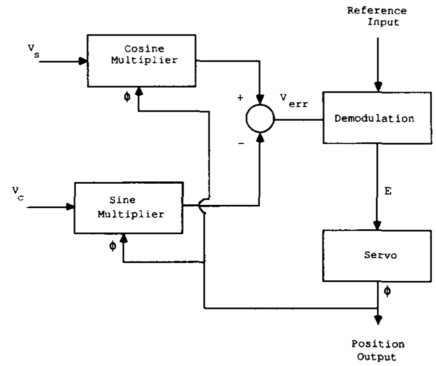

Fig. 2. Resolver-to-digital converter block diagram.

C. Ideal Tracking R/D Conversion

Given the operational block diagram of a tracking R/D converter as shown in Fig. 2, the ideal operation of an R/D converter is described as follows [1], [2]. First, the inputs to the R/D converter, (2), are multiplied by $\cos(\phi)$ and $\sin(\phi)$ respectively where ϕ is the converter's current estimate of the actual position θ. Second, the difference between the two signals is found, giving the error voltage

$$V_{err} = V_1\sin\theta\cos\phi\cos\omega t - V_1\cos\theta\sin\phi\cos\omega t.$$
$$= V_1\sin(\theta - \phi)\cos\omega t. \qquad (3)$$

This double sideband modulated signal is synchronously demodulated, using the resolver excitation signal as a demodulation reference, leaving the envelope signal

$$E = V_1\sin(\theta - \phi). \qquad (4)$$

Finally, this signal is driven to zero by a type II servo loop with ϕ being the control variable. When this is done, the converter's position estimate ϕ is forced to equal the actual shaft position θ resulting in zero position error. In an actual R/D converter, ϕ is continuously updated and output in digital format.

Since the servo loop forms a type II control system, the converter exhibits zero error both when the resolver is at rest and during constant velocity rotation. Moreover, during periods of acceleration the difference between θ and ϕ remains finite and no position error accumulates unless converter tracking rate is exceeded. Hence, after acceleration returns to zero, the converter will again exhibit zero position error.

III. Nonideal Resolver Signals and Their Effect on Position Accuracy

There are numerous ways that the signals generated by a resolver can differ from the ideal situation described above. The most commonly occurring nonideal characteristics are: A) amplitude imbalance, B) quadrature error, C) inductance harmonics, D) reference phase shift, E) excitation signal distortion, and F) disturbance signals. In the following, each of these nonideal characteristics are considered separately. For each, an expression is found relating position error to the degree of nonideal characteristic present under the assumption that the R/D conversion is ideal as described above.

A. Amplitude Imbalance

Amplitude imbalance between the two resolver signals is the most prevalent nonideal resolver signal characteristic. It results from either unbalanced excitation of the resolver phases or from the existence of unequal inductances in the phases. To determine the position error created by amplitude imbalance, let the resolver output signals, given by (2) in the ideal case, be

$$V_s = V_1 \sin\theta \cos\omega t$$
$$V_c = (1+\alpha) V_1 \cos\theta \cos\omega t \tag{5}$$

where α represents the amount of imbalance between the two signals. Following the ideal R/D conversion process outlined earlier, these two signals are processed by a tracking R/D converter. The envelope signal remaining after demodulation can readily be shown to be

$$E = V_1\{\sin(\theta - \phi) - \alpha \cos\theta \sin\phi\}. \tag{6}$$

Driving this signal to zero, as commanded by the servo loop, does not lead to the desired solution $\theta = \phi$. However, by setting the above equation to zero, it is possible to find the position error $\epsilon = \theta - \phi$ that results. Doing so, one obtains

$$\epsilon = \sin^{-1}\left\{\frac{\alpha}{\alpha+2}\sin(\theta+\phi)\right\}. \tag{7}$$

For the realistic case when α is small, the position error is also small which implies that $\sin\epsilon \approx \epsilon$ and $\theta + \phi \approx 2\theta$. Using these approximations, (7) becomes

$$\epsilon \approx \frac{\alpha}{2}\sin(2\theta). \tag{8}$$

This expression shows that the output of an R/D converter will exhibit a position error that is sinusoidal with respect to position, has a frequency twice that of the resolver signals themselves, and has a peak amplitude that is one half the amplitude of the signal mismatch.

It is important to note that the small position error approximation used to obtain (8) is very reasonable, since the acceptable position error in applications using an R/D converter would unlikely be larger than 2^{-10}th of a period (6.13e-3 rad), which is the minimum resolution of commercially available R/D converters.

B. Imperfect Quadrature

Imperfect quadrature between the two resolver signals refers to the situation when the two resolver inductances are not exactly $\pi/2$ radians out of phase with each other. This occurs when the two resolver phases are not machined or assembled in perfect spatial quadrature. To model this situation, let the resolver output signals be

$$V_s = V_1 \sin\theta \cos\omega t$$
$$V_c = V_1 \cos(\theta+\beta)\cos\omega t \tag{9}$$

where β represents the amount of imperfect quadrature. Applying these voltages to an R/D converter, the signal that remains after demodulation is

$$E = V_1\{\sin\theta\cos\phi - \cos(\theta+\beta)\sin\phi\}. \tag{10}$$

As before, the servo loop drives this signal to zero by controlling ϕ. Doing so mathematically, and letting $\cos\beta \approx 1$ and $\sin\beta \approx \beta$, which implies that β will be small in actual applications, one obtains

$$\sin\epsilon \approx -\beta \sin\theta \sin\phi. \tag{11}$$

For small β, the position error $\epsilon = \theta - \phi$ is also small allowing (11) to be further simplified as

$$\epsilon \approx -\beta \sin^2\theta = -\frac{\beta}{2}(1 - \cos 2\theta). \tag{12}$$

This expression relates position error to the degree of imperfect quadrature that exists betwen the resolver signals. Like the error due to amplitude imbalance, this error is periodic with respect to position. However, in this case, the mean error is nonzero and the peak error is equal to the quadrature error.

C. Inductive Harmonics

In general it is impossible to construct a resolver with inductance profiles that are perfect sinusoidal and cosinusoidal functions of position. More generally, the inductances will contain harmonics, and in the case of VR resolvers, will have a dc component. Thus, the mutual inductances can be written in terms of a Fourier series as

$$M_s(\theta) = M_0 + \sum_{n=1}^{\infty} M_n \sin(n\theta)$$
$$M_c(\theta) = M_0 + \sum_{n=1}^{\infty} M_n \cos(n\theta). \tag{13}$$

Substituting (13) into (1) and performing operations that led to (2), the resolver output signals are

$$V_s = V_1\left\{K_0 + \sum_{n=1}^{\infty} K_n \sin(n\theta)\right\}\cos\omega t$$
$$V_c = V_1\left\{K_0 + \sum_{n=1}^{\infty} K_n \cos(n\theta)\right\}\cos\omega t \tag{14}$$

where $K_n = M_n/M_1$. Applying these signals to an R/D converter the envelope signal remaining after demodulation is

$$E = \sqrt{2}K_0\cos(\phi + \pi/4) + \sum_{n=1}^{\infty} K_n \sin(n\theta - \phi). \tag{15}$$

Setting this expression to zero, as is commanded by the servo system, it is once again possible to approximate the position error ϵ. To do so, it is necessary to write the argument of the sine term in (15) as $[(n-1)\theta + \epsilon]$ and to use the trigonometric identity for the sine of the sum of two angles. Performing these steps, (15) simplifies to

$$0 = \sqrt{2}K_0\cos(\phi+\pi/4) + \sum_{n=1}^{\infty} K_n \cdot\left[\sin((n-1)\theta)\cos\epsilon + \cos((n-1)\theta)\sin\epsilon\right]. \tag{16}$$

It is not possible to obtain an exact expression for the position error that is the solution to (16) when the dc term in the Fourier series K_0 is nonzero. However, when the harmonic amplitudes are small, i.e., $K_n \ll 1$ for $n > 1$, ϵ will be small also, which implies that $\cos \epsilon \approx 1$ and $\sin \epsilon \approx \epsilon$. Using these approximations, (16) simplifies to

$$0 \approx \sqrt{2} K_0 \cos(\theta + \pi/4) + \sum_{n=1}^{\infty} K_n \sin((n-1)\theta) + \epsilon \sum_{n=1}^{\infty} K_n \cos((n-1)\theta). \quad (17)$$

The first term in the last summation in (17) is equal to one and all remaining terms are small because of the assumption made above. Because of this, the last summation can be approximated as one and the position error that results from the existence of inductance harmonics can be approximated as

$$\epsilon \approx -\sqrt{2} K_0 \cos(\theta + \pi/4) - \sum_{n=1}^{\infty} K_n \sin((n-1)\theta). \quad (18)$$

This expression illuminates several key ideas regarding the position error incurred due to inductance harmonics. They are: a) the error is proportional to the ratio of the inductive harmonics to the amplitude of the desired fundamental component, b) the error sensitivity due to the dc term is greater than that due to any single harmonic, and c) by comparing the summation in (18) to $M_s(\theta)$ in (13) it is apparent that the general shape of inductive harmonics is preserved in the error expression with a downward shift in frequency. That is, the (n)th inductance harmonic determines the amplitude of the $(n-1)$th harmonic of the position error.

D. Reference Phase Shift

In the discussion of ideal resolver operation, the speed voltage term was ignored since it only exists under dynamic conditions. When resolver velocity is not zero, this term does degrade position accuracy if the demodulation of the resolver signals is not perfectly synchronous. In reality, exact synchronous demodulation does not occur because the resistance of the resolver excitation windings causes a phase shift between the resolver output signals and the resolver excitation signal that is used as the demodulation reference.

To determine the position eror incurred due to this phase shift, reconsider the application of the ideal resolver signals as given by (2), including the speed voltage term, to an R/D converter. Performing the operations leading to the signal to be demodulated gives

$$V_{err} = V_1 \sin(\theta - \phi) \cos \omega t + \frac{\dot{\theta}}{\omega} V_1 \cos(\theta - \phi) \sin \omega t. \quad (19)$$

Synchronous demodulation of this signal is accomplished mathematically by multiplying it by the reference signal $2\cos \omega t$, and filtering all components not translated to around zero frequency. To model nonideal demodulation, (19) is multiplied by $2\cos(\omega t + \delta)$ where δ is the phase shift between the resolver output signals and that of the demodulation reference. Using this reference signal, the envelope signal remaining after demodulation is

$$E = V_1 \left\{ \sin(\theta - \phi) \cos \delta - \frac{\dot{\theta}}{\omega} \cos(\theta - \phi) \sin \delta \right\}. \quad (20)$$

Comparing this equation to the ideal case given by (4), it is apparent that as reference phase shift increases, the amplitude of the desired signal decreases and the amplitude of the undesired speed voltage term increases. Setting this signal equal to zero and solving for the resulting position error gives

$$\epsilon = \tan^{-1}\left\{ \frac{\dot{\theta}}{\omega} \tan \delta \right\} \approx \frac{\dot{\theta}}{\omega} \delta \quad (21)$$

where the second form holds when ϵ and δ are small. The position error in this case is a constant with respect to position and is proportional to the ratoio of the resolver shaft speed to the frequency of the resolver excitation. Thus, in general, it is beneficial to use a high resolver excitation frequency.

E. Excitation Signal Distortion

In all of the preceding analysis it was assumed that the resolver excitation signal was an ideal cosinusoid and contained no additional harmonics. In general, the excitation signal does contain harmonics and thus the resolver signals, given by (2) in the ideal case, are more generally written as

$$V_s = \sin\theta \sum_{n=1}^{\infty} V_{sn} \cos(n\omega t)$$
$$V_c = \cos\theta \sum_{n=1}^{\infty} V_{cn} \cos(n\omega t). \quad (22)$$

Also, it has been assumed that the demodulation waveform is a simple cosinusoid, when in fact the waveform more commonly resembles a square wave. Thus, the demodulation waveform is more appropriately described by the Fourier series

$$V_d = \sum_{n=1}^{\infty} D_n \cos(n\omega t). \quad (23)$$

Applying the resolver signals, (22), to an R/D converter, the error voltage to be demodulated is

$$V_{err} = \sin\theta \cos\phi \sum_{n=1}^{\infty} V_{sn} \cos(n\omega t) - \cos\theta \sin\phi \sum_{n=1}^{\infty} V_{cn} \cos(n\omega t). \quad (24)$$

Demodulating this signal by multiplying by (23) and filtering all components not translated to around zero frequency gives

$$E = \frac{1}{2} \sin\theta \cos\phi \sum_{n=1}^{\infty} V_{sn} D_n - \frac{1}{2} \cos\theta \sin\phi \sum_{n=1}^{\infty} V_{cn} D_n. \quad (25)$$

Finally, driving this signal to zero and assuming that ϵ is small, the resulting position error is

$$\epsilon \approx \frac{\sum_{n=1}^{\infty} (V_{cn} - V_{sn})}{\sum_{n=1}^{\infty} (V_{cn} + V_{sn})} \sin 2\theta. \tag{26}$$

Clearly, if the resolver excitation signals have identical harmonic content, the numerator of (26) is zero and no position error is incurred. However, in the case when the harmonic content is different in each phase, the position error incurred has the same functional shape versus position as that for amplitude imbalance. In fact, (26) is just a generalization of the amplitude imbalance result derived earlier in (8).

F. Disturbance Signals

In applications, an R/D converter is often placed some distance from the resolver, and thus the interconnection between the two is susceptible to disturbance signals that may affect position accuracy. To model this phenomenon, let the resolver output signals be written as

$$V_s = V_1 \sin\theta \cos\omega t + n_s(t)$$
$$V_c = V_1 \cos\theta \cos\omega t + n_c(t), \tag{27}$$

where $n_s(t)$ and $n_c(t)$ are additive disturbance signals. Processing these signals, the error signal remaining before demodulation is

$$V_{\text{err}} = V_1 \sin(\theta - \phi)\cos\omega t + n_s(t)\cos\phi - n_c(t)\sin\phi. \tag{28}$$

Demodulation of this signal, using the general demodulation signal given by (23) gives

$$E = V_1 \sin(\theta - \phi) + \left[n_s(t)\cos\phi - n_c(t)\sin\phi\right] \cdot \sum_{n=1}^{\infty} D_n \cos(n\omega t) \tag{29}$$

where the filtering of components above the excitation frequency has been done to the first term but not to the second.

The frequency spectrum of this second term is a function of the spectrums of the disturbance signals. If the disturbance signals are dc voltages due to amplifier offset voltages for example, the second term in (29) is discarded by demodulation filtering and no position error is incurred. If the disturbance signals contain frequency components at or near any of the demodulation signal harmonics, then these components are aliased down to dc and add directly to the first term in (29). This leads to significant position error as the servo loop finds the angle ϕ that drives (29) to zero. In environments where significant high frequency noise is present, it is possible to low-pass filter the R/D converter input signals or, on some converters, the signal remaining before demodulation V_{err}. The amount of filtering is limited by the inherent phase shift exhibited by the filter at the reference frequency, as this phase shift adds to any reference phase shift, which is itself a source of position error.

TABLE I
SUMMARY OF ERROR SOURCES AND ERROR EXPRESSIONS

Position Error Source	Error Expression	Parameter Description
Amplitude Imbalance	$\frac{\alpha}{2}\sin(2\theta)$	α = additive amplitude difference
Imperfect Quadrature	$-\frac{\beta}{2}(1-\cos 2\theta)$	β = angular deviation from perfect quadrature
Inductive Harmonics	$-\sqrt{2}K_0\cos(\theta+\pi/4) - \sum_{n=1}^{\infty} K_n \sin((n-1)\theta)$	$K_n = \frac{(n)\text{th harmonic}}{\text{fundamental}}$
Reference Phase Shift	$\frac{\theta}{\omega}\delta$	δ = demodulation phase difference
Excitation Distortion	$\frac{\sum_{n=1}^{\infty}(V_{cn} - V_{sn})}{\sum_{n=1}^{\infty}(V_{cn} + V_{sn})}\sin 2\theta$	V_{sn}, V_{cn} = harmonic amplitudes

IV. POSITION ERROR ANALYSIS

The position error resulting from the first five nonideal characteristics considered above are summarized in Table I. An actual resolver and R/D converter position sensing system will be subject to some mixture of all the error sources listed in Table I. As a result, it is natural to ask how each nonideal characteristic contributes to the total position error in the system. For the case when all nonideal characteristics are small, or more rigorously in the limit as all nonideal characteristics approach zero, the overall position error is equal to the sum of the individual position errors. Since the peak errors associated with the nonideal characteristics do not necessarily coincide with each other, it is reasonable and conservative to approximate the overall peak position error as the sum of the individual peak position errors.

To get an idea of the magnitude of each nonideal characteristic allowable for a specific position accuracy degradation it is necessary to define a practical error measure. R/D converters typically have resolutions of 10–16 b per period. If a one-half least significant bit (1/2 LSB) accuracy degradation is considered allowable, then the maximum allowable position error is

$$|\epsilon|_{\max} = \pi(2^{-n}) \tag{31}$$

where n is the number of bits of accuracy desired. Therefore, a 1/2 LSB accuracy degradation out of 10 b (16 b) corresponds to a maximum allowable position error of 3.1E-3 rad (4.8E-5 rad).

In the following paragraphs, bounds on the nonideal resolver characteristics that provide accuracy degradations of 1/2 LSB out of 10 and 16 b are determined.

Given (8), the amplitude imbalance allowed to achieve a specified minimum accuracy is equal to twice the value computed in (31). Thus, to exhibit a maximum 1/2 LSB error out of 10 b (16 b), the amplitude imbalance allowed between the two resolver signals is 0.62% (0.0096%).

The peak position error due to imperfect quadrature is equal to the magnitude of the quadrature error itself. So, for 1/2 LSB error out of 10 b (16 b), the two resolver signals must be within 3.1E-3 rad (4.8E-5 rad) of each other.

When considering inductive harmonics it is convenient to consider the existence of a dc component independently from the existence of higher harmonics. With reference to the first term in (18), the dc inductive component must be no larger than 0.22% (0.0034%) of the amplitude of the desired fundamental inductive component for a maximum 1/2 LSB error out of 10 b (16 b).

Given the second term in (18), there are several ways to evaluate the bounds on the higher inductive harmonics. In the simple, but unrealistic case, when only one undesired harmonic exists, a 1/2 LSB error out of 10 b (16 b) allows the sole harmonic amplitude to be at most 0.31% (0.0048%) the amplitude of the fundamental component.

If the resolver inductance profile has a continuous derivative, then the amplitudes of the higher harmonics fall off at least as fast as $1/n^2$ [10]. Thus, under the assumption that $K_n = k/n^2$ for $n > 1$, an upper bound on the error due to inductive harmonics is

$$|\epsilon|_{\max} < k \sum_{n=2}^{\infty} \frac{1}{n^2} = k\left(\frac{\pi^2}{6} - 1\right). \tag{32}$$

From (32), the maximum allowable k for a maximum 1/2 LSB error out of 10 b (16 b) is 4.7E-3 (7.4E-5). This implies that the second harmonic of the resolver inductance can be no more than 0.12% (0.0018%) of the amplitude of the fundamental component.

Using the second form given in (21) to describe the position error due to reference phase shift, and assuming a worst case scenario where the rotational speed equals the excitation frequency, the maximum phase difference allowed between the resolver output signals and the demodulation reference is equal to the phase shift itself. So, for 1/2 LSB error out of 10 b (16 b) the maximum allowable reference phase shift is 3.1E-3 rad (4.8E-5 rad).

To summarize, the above numerical results are shown in Table II.

TABLE II
SUMMARY OF ERROR MAGNITUDES FOR MAXIMUM 1/2 LSB ERROR

Position Error Source	Error Bound for 1/2 LSB Accuracy Degradation out of:	
	10 bits	16 bits
Amplitude Imbalance	0.62 %	0.0096 %
Imperfect Quadrature	3.1E-3 rad	4.8E-5 rad
DC Inductive Component	0.22 %	0.0034 %
Single Harmonic	0.31 %	0.0048 %
Reference Phase Shift	3.1E-3 rad	4.8E-5 rad

V. CONCLUSIONS

Given the relatively small error bounds given in Table II, it is clear that while R/D converters are capable of high position accuracy, it is difficult to construct a resolver that is capable of generating signals of sufficient quality to exploit this capability. Of the error sources considered, amplitude imbalance, reference phase shift, excitation signal distortion, and disturbance signals are relatively easy to minimize, while the remaining error sources, quadrature error and inductive harmonics, are difficult to minimize.

The former error sources can be minimized through careful engineering practice and design. Amplitude imbalance can be corrected by attenuating the larger of the two signals. If dynamic position accuracy is important, reference phase shift is easily minimized by placing an appropriate phase shift network before the reference input to the R/D converter. Since position error is incurred only when there is a difference between the excitation signals in the two resolver phases, it only necessary to make sure that all routing and processing of the two resolver signals is handled identically to minimize this source of position error. Disturbance signals corrupting the resolver signals are likewise minimized by careful signal routing, shielding and possibly filtering.

The quadrature error and inductive harmonics error sources are difficult to minimize through careful engineering practice and design. The way in which these error sources affect the resolver signals does not lead to straightforward methods to minimize them. Future research will focus on signal processing techniques to minimize these sources of error.

REFERENCES

[1] G. Boyes, Ed., *Synchro and Resolver Conversion.* Analog Devices Inc., 1980.
[2] *Synchro Conversion Handbook.* ILC Data Device Corporation, 1982.
[3] J. Sylvan, "Resolver-to-digital IC makes move to monolithic," *Electron. Design*, pp. 84-91, Feb. 5, 1987.
[4] V. Del Toro, *Electric Machines and Power Systems.* Englewood Cliffs, NJ: Prentice-Hall, 1985.
[5] H. Martinez, "Measuring the position of mechanical shafts," *Power Conv. Int.*, vol. 9, no. 9, pp. 39-42, Oct. 1983.
[6] D. C. Hanselman, R. E. Thibodeau, and D. J. Smith, "Variable-reluctance resolver design guidelines," in *Proc. IECON'89*, (Philadelphia PA), 1989, pp. 203-208.
[7] K. A. Nagar and R. E. Lordo, "Reluctance type resolver with unwound rotor and harmonically graded airgap," in *Proc. Motor-Con '86*, 1986, pp. 27-31.
[8] I. Cushing, "A new high accuracy angular position transducer," in *Proc. Motor-Con '84* (Atlantic City, NJ), 1984.
[9] J. Pavlet and T. J. Harned, "High acceleration, two-phase brushless servo," *Power Conver. Intell. Motion*, vol. 14, no. 1, pp. 20-25, Jan. 1988.
[10] C. McGillem and G. Cooper, *Continuous and Discrete Signal and System Analysis.* New York: Holt, Rinehart and Winston, 1984.