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On the Robustness of $R$-$2R$ Ladder DAC’s

Michael Peter Kennedy, Fellow, IEEE

Abstract—A model of the linear $R$-$2R$ ladder digital-to-analog converter (DAC) is developed in terms of the ratios of the effective resistances at the nodes of the ladder. This formulation demonstrates clearly why an infinite number of different sets of resistors can produce the same linearity error and shows how this error can be reduced by trimming. The relationship between the weights of the bits and the resistor ratios suggests appropriate trimming, design, and test strategies.

Index Terms—Data converters, digital-to-analog conversion, mixed-signal circuits, resistive ladders.

I. INTRODUCTION

MUCH theoretical work in recent years has been devoted to the problem of testing analog and mixed-signal integrated circuits [1]–[8]. In particular, the element-value solvability problem [8] is concerned with determining whether or not it is possible to find the values of (possibly faulty) parameters of a circuit from a set of measurements. This is related to the problem of selecting a limited number of testpoints to perform a test efficiently using a minimum number of measurements [8].

The majority of the circuit theoretic studies of fault location and element solvability assume that a test engineer has access to a sufficiently large number of nodes in the circuit under test. While this may be a valid assumption for board-level work, it does not hold for many integrated circuits, where a limited number of variables may be accessible. An extreme case is a data converter where a single input or output is available. Here, a nonunique relationship between element values and linearity error can produce robustness of the functionality against variations in internal parameter values.

It is well known that the linearity error of an $R$-$2R$ ladder DAC may be reduced by trimming the resistors appropriately. What may appear surprising is that a given trimming procedure can improve the linearity of the device by moving the resistors away from their nominal values. This property results from the structural robustness of the ladder.

In this work, we derive a simplified model, in terms of resistance ratios, of a digital-to-analog converter (DAC) based on a resistive ladder [9] which consists of linear resistors and ideal open/short switches. We study the connection between the weights of the bits and the resistance ratios in order to gain insight into the robustness of the resistive ladder architecture.

In particular, we show that given access to all digital inputs of the DAC, and to only one output node, it is impossible to determine the values of the resistors in the ladder. Only resistor ratios are important in determining the transfer characteristic of the DAC, and these ratios can be determined in principle from a limited set of measurements. This observation can be exploited in defining model-based trim, design, and test strategies [10] for $R$-$2R$ ladder DAC’s.

II. THE MODEL

Throughout this work, we consider the $N$-bit $R$-$2R$ ladder shown in Fig. 1. We extend our analysis in Section V to include also a segmented resistive ladder architecture.

Associated with each node $k$ of the ladder is a pair of linear resistors, $R_{k,1}$ and $R_{k,2}$, which connect it to nodes $k-1$ and $k'$, respectively. An ideal open/short switch connects node $k'$ to the OUT₀ or OUT₁ node, depending on whether the corresponding input bit $b_k$ is 0 or 1.

For notational convenience, we denote by $R_{k,3}$ the effective resistance at node $k$ seen looking into the left-hand end of $R_{k,1}$. In addition, we define the ratios

$$r_k = \frac{R_{k,3}}{R_{k,2}}, \quad k = 1, 2, \ldots.$$ (1)

The $R$-$2R$ ladder is typically used in one of two ways to construct a DAC. Current mode exploits current division along the ladder while voltage mode is based on voltage division [11]. In this work, we treat only voltage-mode operation.
Fig. 3. (a) Equivalent circuit for calculating the contribution $E_{N_i,N_i-1}$ of $V_{N_i-1'}$ to $V_{\text{OUT}}$. (b) Its simplified Thévenin equivalent.

III. VOLTAGE-MODE OPERATION

A DAC exploiting an $R$-$2R$ ladder in voltage mode is shown in Fig. 2. In this case, bit $k_i$ of the input word causes node $k_i'$ to be connected to ground or to $V_{\text{IN}}$ if $k_i = 0$ or 1, respectively.

Since the $R$-$2R$ ladder we consider is linear, the superposition theorem [12] applies, and the voltage at the output node $N$ may be determined by summing the contributions from each of the inputs $V_{k_i'}$ with all other sources zeroed. Thus

$$V_{\text{OUT}} = \sum_{k=1}^{N} E_{N_i,k}$$

where $E_{N_i,k}$ is the contribution to the voltage at node $N$ due to voltage $V_{k_i'}$ applied at node $k_i'$.

Consider first the contribution due to $V_{N_i'}$ with all other sources zeroed. In this case, node $N$ is connected to $V_{N_i'}$ via $R_{N_i,2}$ and to ground via the equivalent resistance $R_{N_i,3}$. By voltage division

$$E_{N_i,N_i} = \frac{R_{N_i,3}}{R_{N_i,2} + R_{N_i,3}} V_{N_i'} \left( R_{N_i,2} \right) \frac{r_N}{1 + r_N} V_{N_i'}$$

where $r_N$ is as defined in (1).

At node $N - 1$, the equivalent circuit for calculating the contribution to $V_{\text{OUT}}$ of $V_{N_i-1'}$ acting alone is shown in Fig. 3(a).

TABLE I

<table>
<thead>
<tr>
<th>k</th>
<th>$R_{k,1}(\Omega)$</th>
<th>$R_{k,3}(\Omega)$</th>
<th>$r_k$</th>
<th>$w_k$</th>
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TABLE II

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<th>$R_{k,3}(\Omega)$</th>
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Fig. 4. Linearity error associated with the ladders detailed in Tables I and II.

$R_{N-1,3}$ denotes the total resistance seen by node $N-1$ looking into $R_{N-1,1}$.

The contribution due to $V_{N-1'}$ acting alone is

$$E_{N_i,N_i-1} = \frac{R_{N_i,3}}{R_{N_i,2} + R_{N_i,3}} \cdot \frac{r_N}{1 + r_N} \cdot V_{N_i-1'}$$

Repeating this process along the ladder, it can be shown in general that $E_{N_i,k} = w_k V_{k_i'}$ for $k = 1, 2, \ldots, N$, where

$$w_k = \begin{cases} \frac{r_N}{1 + r_N} & \text{if } k = N \\ \left( \frac{r_k}{1 + r_k} \right) \prod_{j=k+1}^{N} \left( \frac{1}{1 + r_j} \right) & \text{if } k < N. \end{cases}$$

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The voltage applied at node $k'$ is 0 or $V_{IN}$, depending on whether $b_k=0$ or 1. Thus, $V_{k'} = b_kV_{IN}$. The total output voltage is given by

$$V_{OUT} = \sum_{k=1}^{N} E_{N,k}$$

$$= \sum_{k=1}^{N} b_kw_kV_{IN}$$

$$= [b_1 \ b_2 \ \cdots \ b_k \ \cdots \ b_{N-1} \ b_N] [w_1 \ w_2 \ \cdots \ w_k \ \cdots \ w_{N-1} \ w_N] V_{IN}.$$ 

A. Operation of the Ideal $R$-$2R$ Ladder

In an ideal $R$-$2R$ ladder, $R_{k=1} = R_{k=2} = k = 1, 2, \cdots, N$. Hence, $r_k = 1$ for $k = 1, 2, \cdots, N$ and $w_k = 1/2^{N-k+1}$. Therefore

$$V_{OUT} = \sum_{j=1}^{N} \frac{b_j}{2^{N-k+1}} V_{IN}$$

$$= U \frac{V_{IN}}{2^N}$$

where $b_Nb_{N-1} \cdots b_2b_1$ is the binary expansion of the input word $U$.

B. Determination of Resistance Ratios

We ask the question: can one determine the ratios $r_k$, $k = 1, 2, \cdots, N$ in an $R$-$2R$ ladder DAC simply by measuring the output voltage $V_{OUT}$?

Let $V_{OUT}(U)$ be the measured output corresponding to input word $U$, as before. In the voltage-mode case, a judiciously chosen subset of $N$ measurements (out of a possible $2^N$) is sufficient to determine the weights $w_k$. In particular, we have that

$$\begin{bmatrix} V_{OUT}(1) \\ V_{OUT}(2) \\ \vdots \\ V_{OUT}(2^{N-2}) \\ V_{OUT}(2^{N-1}) \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & \cdots & 0 & 0 \\ 0 & 1 & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & 0 & \cdots & 1 & 0 \\ 0 & 0 & 0 & \cdots & 0 & 1 \end{bmatrix} \begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_{N-1} \\ w_N \end{bmatrix}$$

(3)

where $V_{OUT}$ is measured at $U = 1, 2, 4, 8, \cdots, 2^N$. Thus, the weights $w_k$ of a voltage-mode DAC can in principle be determined with just $N$ measurements, provided that $V_{IN}$ is known. From these weights $w_k$, the ratios $r_k$ may be estimated by setting

$$\hat{r}_N = \frac{w_N}{1-w_N}$$

and evaluating

$$\hat{r}_k = \frac{\rho_k}{1-\rho_k}$$

for $k = (N-1)$ to 1 in turn, where

$$\rho_k = w_k \prod_{j=k+1}^{N} (1 + \hat{r}_j).$$
Now $\hat{r}_k$ provides an estimate of $r_k$. In a process monitoring role, these estimates could potentially be used to quantify the deviation of production parts from their nominal design values. From a test engineering perspective, the extracted weights $w_k$ can be exploited in linear error mechanism modeling [8], [10].

C. Example

Consider the two voltage-mode R-2R ladder DAC’s whose resistor values are given in Tables I and II, respectively. Here, the ladders are mismatched in a similar way but the normalized resistances of the ladders are different (10 and 20 kΩ, respectively). Output measurements are simulated for $V_{IN} = 5$ V in both cases. While the values of the resistors in the 20 kΩ ladder are not quite double those in the 10 kΩ ladder, they have been chosen so that the ratios $r_k$ and weights $w_k$ are identical. Therefore, the normalized error plots for these devices, shown in Fig. 4, are also identical; equivalently, both devices belong to the same ambiguity group [7], [8].

In both of these examples, the estimates $\hat{r}_k$, $k = 1, 2, \cdots, 8$ of the resistance ratios determined from simulations of the two ladders are identical to ten decimal places.

IV. RELATIONSHIP BETWEEN WEIGHTS $w_k$ AND RATIOS $r_k$

It is interesting to note the form of the weights in the eight-bit case

$$w_1 = \frac{r_1}{1+r_1} \frac{1}{1+r_2} \frac{1}{1+r_3} \frac{1}{1+r_4} \frac{1}{1+r_5} \frac{1}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_2 = \frac{r_2}{1+r_2} \frac{1}{1+r_3} \frac{1}{1+r_4} \frac{1}{1+r_5} \frac{1}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_3 = \frac{r_3}{1+r_3} \frac{1}{1+r_4} \frac{1}{1+r_5} \frac{1}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_4 = \frac{r_4}{1+r_4} \frac{1}{1+r_5} \frac{1}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_5 = \frac{r_5}{1+r_5} \frac{1}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_6 = \frac{r_6}{1+r_6} \frac{1}{1+r_7} \frac{1}{1+r_8}$$

$$w_7 = \frac{r_7}{1+r_7} \frac{1}{1+r_8}$$

$$w_8 = \frac{r_8}{1+r_8} .$$

$^1$ $R_{x,1}$ and $R_{x,2}$ in Table II have been “trimmed” to compensate for the error in $R_{x,2}$.

A. Implications for Trimming

In an ideal binary-weighted DAC, we require that $w_{k+1} = 2w_k$ for all $k$. This can be achieved by ensuring that

$$r_{k+1} = \frac{2r_k}{1+r_k} .$$

In a nominal R-2R ladder, $r_k = 1$ for all $k$. If, due to production variations, $r_k \neq 1$ for some $k$, the constraint (4) can still be met, and the linearity error minimized, by adjusting $r_j$ for $j = k + 1, k + 2, \cdots, N$. Each $r_j$ can be set by trimming $R_{x,1}$ and/or $R_{x,2}$. Note that, during the trimming process, it may be necessary to move resistors away from their nominal values.

If the ladder is trimmed from the right end by adjusting the ratios $r_1, r_2, r_3$, etc., in turn, it is clear that the absolute value of each weight $w_k$ will be affected by an adjustment of $r_j$ for all $j > k$. However, the ratio of any pair of weights $(w_i/w_j \ i < k)$ is unaffected by trimming further up the ladder. Therefore, the trimming algorithm should try to fix the ratios of weights with the current value of the LSB rather than its final value.

B. Implications for Design

From the designer’s perspective, the goal is to ensure that $w_{k+1} = 2w_k$ in order to produce a binary-weighted DAC. Clearly, this objective can be achieved with any number of different sets of ratios $r_k$. In particular, it is not necessary to choose $r_k = 1$, nor is it necessary to define the absolute value of $w_1$.

Consider the case of an ideal R-2R ladder where we want $r_k = 1$ for all $k$. When switch resistances are taken into ac-
count, a dummy switch can be inserted in series with $R_{1,1}$ to compensate for the switch in series with $R_{1,2}$ and guarantee monotonicity. Alternatively, an appropriate choice of “mismatch” at the right end of the ladder when sizing the switches in series with $R_{1,1}$ and $R_{1,2}$ can yield ratios $r_k \neq 1$ but still guarantee binary weighting. The total switch area resulting from this strategy may be less than by choosing $r_k \equiv 1$ for all $k$.

### C. Implications for Production Monitoring

Finally, from the production monitoring viewpoint, we note that although $w_k$, $k = 1, 2, \ldots, N$, can in principle be determined with just $N$ measurements using (3), a better estimate of the $w_k$’s may be obtained in the case of limited measurement resolution by solving a larger subset of the overdetermined system of equations

$$\begin{bmatrix} V_{\text{OUT}(0)} \\ V_{\text{OUT}(1)} \\ V_{\text{OUT}(2)} \\ \vdots \\ V_{\text{OUT}(2^N-2)} \\ V_{\text{OUT}(2^N-1)} \end{bmatrix} = \begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ 0 & 1 & \cdots & 0 & 0 \\ 0 & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & 0 & 0 \\ 1 & 1 & \cdots & 1 & 1 \end{bmatrix} \begin{bmatrix} w_1 \\ w_2 \\ w_3 \\ \vdots \\ w_N-1 \\ w_N \end{bmatrix}$$

### V. Voltage-Mode Operation: Segmented Architecture

In an $R\cdot 2R$ ladder, it is necessary to have tight matching between each bit and the sum of all lesser bits in order to ensure monotonic operation [13]. Segmented architectures allow this requirement to be relaxed and permit the construction of high-resolution converters.

An $(M + N)$-bit segmented design provides a coarse/fine structure. The most significant $M$ bits define $2^M$ segments which are further subdivided by an $N$-bit $R\cdot 2R$ ladder. Provided that the $N$-bit ladder is monotonic and that its full-scale output is less than that of the next segment, monotonicity is guaranteed. This is called the next-segment approach.

### A. Operation of the Voltage-Mode Segmented $(M + N)$-Bit DAC

A commonly-used next-segment DAC architecture is shown in Fig. 5. The coarse DAC consists of $2^M - 1$ identical resistors ($R_{N+1\cdot 2^M-1,1}$, $R_{N+2\cdot 2^M-1,1}$) which are selected by a thermometer code. The fine DAC is an $N$-bit $R\cdot 2R$ ladder.

The least significant bits are applied directly to the switches in the $R\cdot 2R$ ladder. Bit $b_k$, $k = 1, 2, \ldots, N$, of the input word causes node $k'$ to be connected to ground or to $V_{\text{IN}}$ if $b_k = 0$ or 1, respectively. The most significant $M$ bits are decoded to produce $a_1, a_2, \ldots, a_{2^M-1}$ which select the segments. Bit $a_k$, $k = 1, 2, \ldots, 2^M - 1$ causes node $N + k'$ to be connected to ground or to $V_{\text{IN}}$ if $a_k = 0$ or 1, respectively.

Since this network is linear, the superposition theorem [12] applies, and the voltage at the output node $N + 2^M - 1$ may be determined by summing the contributions from each of the inputs $V_k$ with all other sources zeroed. Thus

$$V_{\text{OUT}} = \sum_{k=1}^{N+2^M-1} E_{N+2^M-1,k}$$

where $E_{N+2^M-1,k}$ is the contribution to the voltage at node $N + 2^M - 1$ due to voltage $V_k$ applied at node $k'$.

Consider first the contribution due to $V_{N+2^M-1}$ with all other sources zeroed. In this case, node $N + 2^M - 1$ is connected to $V_{N+2^M-1}$ via $R_{N+2^M-1,2}$ and to ground via the equivalent resistance $R_{N+2^M-1,2}$.

By voltage division

$$E_{N+2^M-1} = \frac{R_{N+2^M-1,3}}{R_{N+2^M-1,2} + R_{N+2^M-1,3}} V_{N+2^M-1}$$

$$= \frac{\tau_{N+2^M-1} V_{N+2^M-1}}{1 + \tau_{N+2^M-1}}$$

where $\tau_{N+2^M-1}$ is as defined in (1).
The contributions to \( V_{\text{OUT}} \) due to the other inputs may be calculated by determining the Thévenin equivalent to the right of each node in turn, as in the case of the voltage-mode ladder without segmentation. The contribution due to the input \( V_{k'} \) at node \( k' \) is given by

\[
E_{N+2^M-1,k} = w_k V_{k'}
\]

where \( w_k \) is defined by

\[
w_k = \begin{cases} \frac{\tau_N+2^M-1}{1 + \tau_N+2^M-1}, & \text{if } k = N + 2^M - 1 \\ \frac{\tau_k}{1 + \tau_k} \prod_{j=k+1}^{N+2^M-1} \left( \frac{1}{1 + \tau_j} \right), & \text{if } k < N + 2^M - 1. \end{cases}
\]  

(5)

The total output voltage is given by (5a) at the bottom of this page.

**B. Operation of the Ideal Segmented DAC**

In the \( N \)-bit DAC, \( R_{k,3} = R_{k,2} = 2R \) for \( k = 1, 2, \ldots, N \), giving \( \tau_k = 1 \) for \( k = 1, 2, \ldots, N \). The output resistance of the ladder is increased to \( 2R \) by setting \( R_{N+1,2} = R \). The segment resistors have nominal value \( 2R \) and are interconnected by short-circuits. Hence, \( R_{k,2} = 2R \) for \( k = N + 1, N + 2, \ldots, N + 2^M - 1 \) and \( R_{k,1} = 0 \) for \( k = N + 2, N + 3, \ldots, N + 2^M - 1 \). This gives \( \tau_{N+k,3} = 2 \tau_k \) for \( k = 1, 2, \ldots, 2^M - 1 \). Hence

\[ r_{N+k} = \frac{1}{k}, \quad k = 1, 2, \ldots, 2^M - 1. \]

Substituting for each \( r_k \) yields

\[
\begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_{N-1} \\ w_N \end{bmatrix} = \begin{bmatrix} 1 \\ \frac{1}{2^M+1} \\ \frac{1}{2^M+2} \\ \vdots \\ \frac{1}{2^M} \end{bmatrix}
\]

Therefore

\[
V_{\text{OUT}} = \left( \sum_{k=1}^{N} b_k \frac{2^M-1}{2^M-k+1} + \sum_{k=1}^{2^M} c_k \right) \frac{V_{\text{IN}}}{2^M}
\]

where \( b_N b_{N-1} \cdots b_1 b_0 \) are the LSB’s of the input word \( U \) and the upper \( M \) bits are decoded to give the \( c_k \).s.

**C. Diagnosability of the Segmented Voltage-Mode DAC**

Is it possible to determine the ratios \( r_k, k = 1, 2, \ldots, N + 2^M - 1 \) in a segmented voltage-mode DAC simply by measuring the output voltage \( V_{\text{OUT}} \)?

Let \( V_{\text{OUT}}(U) \) be the measured output corresponding to input word \( U \), as before. In this case, the weights \( w_k \) may be determined by making just \( N + M \) (out of a possible \( 2^{N+M} \)) measurements of \( V_{\text{OUT}} \). In particular, \( N + M \) measurements of \( V_{\text{OUT}} \) with \( U = 1, 2, 4, 8, \ldots, 2^N, \ldots, 2^{N+M}, \) yield (6) at the bottom of the next page.

\[
V_{\text{OUT}} = \sum_{k=1}^{N+2^M-1} E_{N,k} = \left( \sum_{k=1}^{N} b_k w_k + \sum_{k=1}^{2^M-1} c_k w_{N+k} \right) V_{\text{IN}}
\]

\[
= \begin{bmatrix} b_1 & b_2 & \cdots & b_k & \cdots & b_{N-1} & b_N & a_1 & a_2 & \cdots & a_{2^M-1} \end{bmatrix} \begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_k \\ \vdots \\ w_{N-1} \\ w_N \\ w_{N+1} \\ w_{N+2} \\ \vdots \\ w_{N+2^M-1} \end{bmatrix} V_{\text{IN}},
\]  

(5a)
Assuming that $V_{IN}$ is known, (6) may be rewritten to give the weights $w_k$ explicitly in terms of the $N + M$ measured outputs shown at the bottom of this page.

From these weights $w_k$, the ratios $r_k$ may be estimated by setting

$$\hat{r}_k = \frac{\rho_k}{1 - \rho_k}$$

and evaluating

$$\hat{r}_k = \frac{\rho_k}{1 - \rho_k}$$

for $k = N + 2^M - 1$ to 1 in turn, where

$$\rho_k = w_k \prod_{j=k+1}^{N+2^M-1} (1 + \hat{r}_j).$$

As before, $\hat{r}_k$ provides an estimate of $r_k$.

D. Example

Consider the 14-bit segmented voltage-mode DAC whose resistor values are given in Table III. The DAC consists of an 11-bit $R-2R$ ladder and three decoded bits driving seven segment resistors. This linear network was simulated using a reference input $V_{IN} = 5$ V. The endpoint-corrected linearity error is shown in Fig. 6. Note that the estimates $\hat{r}_k$ of the resistor ratios

$$\begin{bmatrix}
V_{OUT}(1) \\
V_{OUT}(2) \\
V_{OUT}(4) \\
\vdots \\
V_{OUT}(2^{N-2}) \\
V_{OUT}(2^{N-1}) \\
\vdots \\
V_{OUT}(2^{N+M-1})
\end{bmatrix} =
\begin{bmatrix}
1 & 0 & 0 & \cdots & 0 & 0 & 0 & \cdots & 0 \\
0 & 1 & 0 & \cdots & 0 & 0 & 0 & \cdots & 0 \\
0 & 0 & 1 & \cdots & 0 & 0 & 0 & \cdots & 0 \\
\vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & 0 & 0 & \cdots & 1 & 0 & 0 & \cdots & 0 \\
0 & 0 & 0 & \cdots & 0 & 1 & 0 & \cdots & 0 \\
\vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & 0 & 0 & \cdots & 0 & 1 & 1 & \cdots & 1
\end{bmatrix}
\begin{bmatrix}
w_1 \\
w_2 \\
w_3 \\
\vdots \\
w_{N-1} \\
w_N \\
\vdots \\
w_{N+1} \\
w_{N+2} \\
w_{N+2^M-1}
\end{bmatrix}$$

$$\begin{bmatrix}
w_1 \\
w_2 \\
w_3 \\
\vdots \\
w_{N-1} \\
w_N \\
\vdots \\
w_{N+1} \\
w_{N+2} \\
w_{N+2^M-1}
\end{bmatrix} =
\frac{1}{V_{IN}}
\begin{bmatrix}
V_{OUT}(1) \\
V_{OUT}(2) \\
V_{OUT}(4) \\
\vdots \\
V_{OUT}(2^{N-2}) \\
V_{OUT}(2^{N-1}) \\
\vdots \\
V_{OUT}(2^{N}) \\
V_{OUT}(2^{N+1}) - V_{OUT}(2^{N}) \\
\vdots \\
V_{OUT}(2^{N+M-1}) - \cdots - V_{OUT}(2^{N+1}) - V_{OUT}(2^{N})
\end{bmatrix}.\]
at the node of the network extracted from the simulated data are correct to eight significant figures.

Note also that mismatch errors at the right end of the ladder are less significant than those at the left end or in the segment resistors.

Fig. 7 shows the linearity error of the (3 + 11)-bit DAC summarized in Table IV. In this case, gross errors (of up to 20%) have been introduced into the resistors at the LSB end of the ladder in order to exaggerate the relative effects of mismatches at different bit positions. Here, the large errors in the LSB resistors (and the corresponding weights) contribute proportionately less to the overall error than do the much smaller errors in the segment resistors.

Qualitatively, this is because the full-scale error due to a fractional error in an LSB is smaller than for the same fractional error in a more significant bit. Furthermore, errors in the LSB resistor ratios \( r_k \) are divided down by a factor of approximately four at each node up the ladder so that the ratios for the MSB’s are relatively insensitive to errors further to the right along the ladder.

In terms of trimming a segmented voltage-mode DAC, more effort should be devoted to matching the segment resistors and the MSB’s of the ladder than the LSB’s since the former are more likely to determine the final accuracy than the latter.

VI. CONCLUDING REMARKS

In this work, we have developed a model of the linear \( R-2R \) ladder DAC which is parameterized by the ratios of effective resistances at the nodes of the ladder. This formulation provides insight into the operation of the ladder. It explains why the ladder is insensitive to the absolute values of the constituent resistors, and suggests appropriate trimming, design, and test strategies. We have not considered the case of code-dependent resistors in the ladder [14]; this is the focus of on-going work.

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REFERENCES


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