Distributed Biasing of Differential RF Circuits

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Abstract—A distributed balun is presented as an alternative to inductor-based biasing of active differential circuits. The new scheme has broad bandwidth with low loss and is suited to broad-band multifunctional RF and microwave circuits. Additionally, the distributed biasing circuit discriminates between differential and common-mode signals leading to high common-mode rejection. Measured gains, centered at 580 MHz, of a selected pseudodifferential amplifier using the new biasing circuit confirm the underlying theory.

Index Terms—Balun, broad-band biasing, differential amplifier.

I. INTRODUCTION

Radio-frequency (RF) circuits commonly utilize differential signal paths with the benefit of noise immunity and reduced sensitivity to component variations. On-chip RF power amplifiers are often pseudodifferential where the common current source typically used in fully differential circuits is sacrificed to enable a larger voltage swing (see Fig. 1). Here, the inductors present high RF impedance to the transistors (represented as transconductances), while providing a low-impedance path for bias currents. With a sufficiently high-\(Q\) inductor, RF energy is delivered to the load rather than being dissipated in the bias circuitry. With the relatively low \(Q\) of on-chip planar inductors [1], inductance peaking near self-resonance is utilized and this leads to narrow-band operation. Thus, in critical situations, off-chip inductors are sometimes used in conjunction with on-chip active devices. In addition, inductive biasing of pseudodifferential circuits presents the same environment to common- and differential-mode signals so that the common-mode rejection ratio (CMRR) is one.

This paper presents a biasing scheme for broad-band pseudodifferential circuits leading to high CMRR. The biasing circuitry consists of a coupled-line structure analogous to that of a Marchand balun [2] (see Fig. 2). The balun structure is comprised of transmission lines connected to a pair of coupled lines that normally transforms a single-ended impedance at port 1 to a balanced output at ports 3 and 4. The capacitors in Fig. 2 are purely for miniaturization purposes [3], [4] since the distributed transmission lines are quarter-wavelength long at a frequency well above the passband of the balun. In Section II, the validity of the Marchand balun structure as a biasing circuit is investigated and analysis of the even- and odd-mode impedances of the balanced ports is presented. Practical implementations are presented in Section III.

II. ANALYSIS AND DESIGN

Here, the design equations of a class of distributed differential biasing circuits are developed based on coupled resonator theory. In Section II-A, the equations describing the CMRR are first developed, as maximizing the CMRR in pseudodifferential circuits is a key design objective. The subsequent design equations enable the design of the biasing-circuit network with specified characteristics.
A. Conventional Biasing

Multifunctional systems require broad-band circuits and, for differential circuits, this requires large differential gain $A_d$. At the same time, it is desirable to minimize the common-mode gain $A_c$, as the resulting high CMRR provides immunity to substrate induced noise. With transistor transconductance $g_m$, and total even- and odd-mode impedances $Z_{\text{EVEN}}$ and $Z_{\text{ODD}}$ presented to the drains of the transistors, the gains are approximately

$$A_d = g_m Z_{\text{ODD}} \quad \text{and} \quad A_c = g_m Z_{\text{EVEN}}$$

and thus,

$$\text{CMRR} = \frac{A_d}{A_c} = \frac{Z_{\text{ODD}}}{Z_{\text{EVEN}}}. \quad (2)$$

Thus, the desired amplifier characteristics are obtained by synthesizing the even- and odd-mode impedances.

Modal analysis of the inductor biasing circuit results in Fig. 3, from which the total even-mode impedance is

$$Z_{\text{EVEN}}(s) = \left( sL + \frac{R_{\text{DD}}}{2} \right) / R_L \quad \text{or} \quad \text{(3)}$$

where $s$ is the Laplace operator (\(/\) indicates a parallel connection) and the total odd-mode impedance is

$$Z_{\text{ODD}}(s) = sL / R_L. \quad \text{(4)}$$

Thus, unless the inductance is very large, there will be a strong frequency variation of $Z_{\text{ODD}}$ and $Z_{\text{EVEN}}$. Since $R_{\text{DD}}$ is often negligible, the CMRR is also one. In the differential-mode, the gain is maximized over a broad frequency band for a specific single-ended load impedance $R_L$. However, if the common-mode gain has prescribed frequency characteristics, then the CMRR will be inversely proportional to the even-mode impedance. Thus, low even-mode impedance of the biasing network is desirable, as this suppresses common-mode noise. This defines a vital design objective of the new biasing circuit.

B. New Design Based on Analogy to Coupled Resonators

Coupled resonator structures are appropriate topology choices that present different impedances for the common- and differential-mode signals. In essence, bias is also a common-mode signal and must be accommodated in the chosen topology. A class of appropriate topologies is based on a lumped bandpass resonator coupled into by the inverter $K$, shown in Fig. 4(a). The inverter element is an idealized component that is independent of frequency and can be approximated by a quarter-wavelength-long transmission line, as well as by various other circuit configurations [5]. At resonance, the bandpass resonator presents an open circuit and, following impedance inversion, the input impedance of the network is a short circuit. This is then the kind of network required to be presented to the common-mode signals. The second part of the network synthesis problem is a creation of the high-impedance condition for differential-mode signals. Again, we will consider this issue separately. Fig. 4(b) illustrates a pair of bandpass resonators coupled by inverters. At resonance, the input impedance of the network is infinite. This is because the high impedance of resonator 2 transforms to a short circuit after the second inverter $K_2$, and this subsequently appears as an open circuit at the input of the network. Thus, by adding an extra resonator, another topology suitable for the differential-mode operation is derived. If a load is now connected in parallel with this network, then at resonance, this will be the total impedance presented to the output transistors of the pseudodifferential amplifier. In summary, presenting different impedances for common- and differential-mode signals is desired. It is shown in Section II-C that the Marchand balun structure has the requisite properties.

C. Distributed Biasing

Consider now the topology of the Marchand balun of Fig. 2 with the balanced ports 3 and 4 connected to the output of the pseudodifferential circuit. The aim here is to show that this topology has the requisite even- and odd-mode impedances. Performing modal analysis at the balanced ports, the modal sub-networks of Fig. 5 are obtained. In analyzing this structure, we shall make use of the network model of pair of coupled lines in a homogeneous media [6] (shown in Fig. 6). Direct application of the network model and application of the relevant port conditions, dictated by the modal sub-networks of the biasing circuit at point $X$, simplifies the equivalent sub-networks of Fig. 5 to those of Fig. 7. Application of the relevant Kuroda transformation [5] to distribute the inductance in Fig. 7(c) results in the transformed sub-circuits of Fig. 7(d). Scrutiny in
Fig. 5. Modal sub-networks of the balun-like biasing circuit of Fig. 2 with single-ended load resistance \( R_L \) (assuming negligible power supply source impedance \( R_P \)).

Fig. 6. Pair of symmetrical coupled lines in homogenous media. (a) Physical layout. (b) Equivalent-network model with \( N = (Z_{oc} + Z_{oo}) / (Z_{oc} - Z_{oo}) \).

Fig. 7(b) and (d) reveals that they are, in fact, analogous to the lumped-element coupled resonator examples discussed earlier. The major differences are that the transmission lines replace the inverters and the additional resistor \( R_L \) represents the loss associated with each capacitor. In the distributed domain, each inductor or capacitor also represents a short- or open-circuited stub, respectively, which is a quarter-wavelength long at the frequency \( f_a \). The open-circuited stubs may, however, be approximated by lumped capacitors for miniaturization purposes. Thus, over the operating band, the performance of the even- and odd-mode sub-circuits is very close to their lumped-element counterparts. In summary, centered at a frequency \( f_a \), the structure of the Marchand balun maximizes differential power delivery to the load while minimizing the gain of common-mode signals. DC bias may easily be applied through the short-circuited ports (labeled \( b \) in Fig. 2) with the usage of decoupling capacitors. The mathematical formulations for the total even- and odd-mode impedances with lumped capacitors are derived from Fig. 7(b) and (c) as follows:

\[
Z_{\text{EVEN}}(S) = \frac{N_1 S^2 + N_2 S}{D_1 S^2 + D_2 S + D_3} \tag{5}
\]

with coefficients

\[
N_1 = j (2Z_{oc}^2 Z_1 \pi f C R_L)
\]
\[
N_2 = Z_1 R_L Z_{oc} + Z_{oc}^2 R_L
\]
\[
D_1 = (Z_1 R_L) + j (2Z_{oc}^2 \pi f C Z_1)
\]
\[
D_2 = (Z_{oc}^2 + Z_{oc} Z_1) + j (2Z_{oc} R_L \pi f C Z_1)
\]
\[
D_3 = Z_{oc} R_L
\]

and

\[
Z_{\text{ODD}}(S) = \frac{N_1 S^3 + N_2 S^2 + N_3 S}{D_1 S^3 + D_2 S^2 + D_3 S + D_4} \tag{6}
\]

with coefficients

\[
N_1 = Z_{oc}^2 R_L Z_1
\]
\[
N_2 = j (2Z_{oc}^2 R_L \pi f C Z_1)
\]
\[
N_3 = Z_1 Z_{oc} Z_2 R_L + Z_{oc}^2 Z_2 R_L
\]
\[
D_1 = Z_{oc}^2 Z_1
\]
\[
D_2 = (Z_1 Z_{oc} R_L + Z_{oc} Z_2 R_L) + j (2Z_{oc}^2 \pi f C Z_1)
\]
\[
D_3 = (Z_{oc} Z_2 Z_1 + Z_{oc}^2 Z_2) + j (2Z_{oc} Z_2 \pi f C Z_1 R_L)
\]
\[
D_4 = Z_2 Z_{oc} R_L
\]

In (5) and (6), \( S \) is the Richards transform [5] defined as \( j \tan(\theta) = j \tan((\pi/2)(f/f_a)) \), and \( f_a \) is the frequency at which the distributed lines of the balun are a quarter-wavelength long. In the above equations, \( Z_{oc} \) is the characteristic impedance of the uncoupled transmission lines, while

\[
Z_1 = \frac{\sqrt{Z_{oc} Z_{oo}}}{\sqrt{1 - \left(\frac{Z_{oc} - Z_{oo}}{Z_{oc} + Z_{oo}}\right)^2}} \tag{7}
\]

and

\[
Z_2 = \sqrt{Z_{oc} Z_{oo}} \sqrt{1 - \left(\frac{Z_{oc} - Z_{oo}}{Z_{oc} + Z_{oo}}\right)^2} \tag{8}
\]

with \( Z_{oc} \) and \( Z_{oo} \) being the even- and odd-mode impedances of the coupled lines forming the balun. The total even- and odd-mode impedances presented to the pseudodifferential active circuit are, in general, complex and, thus, may be written in the following notation:

\[
|Z_{\text{EVEN}}(S)| = \sqrt{\Re\{Z_{\text{EVEN}}(S)\}^2 + \Im\{Z_{\text{EVEN}}(S)\}^2} \tag{9}
\]
\[
|Z_{\text{ODD}}(S)| = \sqrt{\Re\{Z_{\text{ODD}}(S)\}^2 + \Im\{Z_{\text{ODD}}(S)\}^2} \tag{10}
\]

Hence, the procedure of determining the circuit parameters of the new biasing circuit is fairly simple. Suitable choices of the characteristic impedances \( Z_1 \) and \( Z_2 \) solve (7) and (8) for the even- and odd-mode characteristic impedances \( Z_{oc} \) and \( Z_{oo} \).
Fig. 7. Equivalent sub-circuits of Fig. 5: (a) utilizing the network model, (b) even-mode sub-circuit, (c) odd-mode sub-circuit, and (d) odd-mode sub-circuit after the Kuroda transformation with \( \frac{1}{2} \frac{1}{90} \frac{0}{61} \frac{1}{49} \frac{1}{61} \frac{90}{0} \frac{41}{90} \frac{90}{41} \frac{309}{252} \).

respectively. Selection of the characteristic line impedance \( Z_{\text{tot}} \) together with choice of capacitance value \( C \) fully determines (5) and (6). The magnitudes of the total even- and odd-mode impedance functions may then be evaluated using (9) and (10). The commensurate length of the transmission lines will be determined by the value of \( C \). Zero capacitance results in the lines being a quarter-wavelength long at the center of the operating band, otherwise miniaturization of the lines is feasible. An example that shows the overall characteristics of the magnitudes of the even- and odd-mode impedances without lumped capacitors is depicted in Fig. 8 for a range of single-ended load resistances \( R_L \).

D. Discriminative Operating Bandwidth

The operating bandwidth of the balun is defined at the crossover frequencies where

\[
|Z_{\text{EVEN}}(S)| = |Z_{\text{ODD}}(S)|.
\]

At the crossover frequency points (see Fig. 8), the CMRR of the pseudodifferential circuit is one and the balun has a similar performance to that of the nondiscriminative inductor-based biasing circuit. For a certain choice of circuit parameters, the frequency points at which the even- and odd-mode impedances crossover may be altered facilitating adjustment of the discriminative bandwidth. This is demonstrated in Fig. 9 through a set of examples using the design procedure described in Section II-C. It is worth emphasizing that the total even- or odd-mode impedance of (5) or (6) is the impedance of the input of the balun in parallel with the single-ended load resistance \( R_L \). This implies that, in order for the pseudodifferential circuit to deliver its power to the load, it is necessary that the output signal at each arm sees the correct impedance. Of course, there will only be a perfect match at the center of the band (since the input balun impedance is then infinite), but not at the crossover frequency points. However, a slight mismatch will only degrade the transfer of power slightly and is, therefore, acceptable. Fig. 8 illustrates this argument for a 50-\( \Omega \) system.
with the lumped capacitors if circuit miniaturization is desired. Modification of the total even- and odd-mode impedances to take into account the effect of substrate, conductor, and capacitor loss $R_c$ leads to

$$Z_{\text{EVEN}}(S) = \frac{N_1 S^2 + N_2 S}{D_1 S^2 + D_2 S + D_3}$$

(12)

with coefficients

$$N_1 = (Z_{\text{ot}}^2 - Z_1 R_L) + j (2Z_{\text{ot}}^2 Z_1 \pi f C R_L R_C)$$
$$N_2 = Z_A R_L Z_{\text{ot}} R_C + Z_{\text{ot}}^2 R_L R_C$$
$$D_1 = (Z_1 R_L R_C + Z_{\text{ot}}^2 Z_1) + j (2Z_{\text{ot}}^2 \pi f C Z_1 R_C)$$
$$D_2 = (Z_{\text{ot}}^2 R_C + Z_{\text{ot}}^2 Z_1 R_C + Z_{\text{ot}} Z_1 R_L) + j (2Z_{\text{ot}} R_L \pi f C Z_1 R_C)$$
$$D_3 = Z_{\text{ot}} R_L R_C$$

and

$$Z_{\text{ODD}}(S) = \frac{N_1 S^3 + N_2 S^2 + N_3 S}{D_1 S^3 + D_2 S^2 + D_3 S + D_4}$$

(13)

with coefficients

$$N_1 = Z_{\text{ot}}^2 R_L Z_1 R_C$$
$$N_2 = (R_L Z_2 Z_{\text{ot}}^2 Z_1) + j (2Z_{\text{ot}}^2 R_L R_C f C Z_1 R_C)$$
$$N_3 = Z_1 Z_2 Z_{\text{ot}} R_L R_C + Z_{\text{ot}}^2 R_L R_C$$
$$D_1 = Z_{\text{ot}}^2 Z_1 R_C$$
$$D_2 = (Z_1 Z_2 R_L R_C + Z_1 Z_{\text{ot}} R_L R_C + Z_{\text{ot}} Z_1 Z_2) + j (2Z_{\text{ot}} Z_2 R_C f C Z_1 R_C)$$
$$D_3 = (Z_2 Z_{\text{ot}} Z_1 R_C + Z_2 Z_{\text{ot}}^2 R_C + Z_1 Z_2 Z_{\text{ot}} R_L) + j (2Z_{\text{ot}} Z_2 R_C f C Z_1 R_C)$$
$$D_4 = Z_2 Z_{\text{ot}} R_L R_C$$

Now $S$ is the general Richards transform [7], defined as

$$\tanh\left(\alpha + j \beta\right) = \tanh\left(\alpha + j \left(\pi / 2\right) \left(f / f_0\right)\right)\text{ and } \alpha$$

is the attenuation constant in nepers/meter. With no miniaturization capacitors (i.e., $C = 0$, $R_c = \infty$) and a finite value of $\alpha$, the same circuit parameters that generated the plots of Fig. 8 are now fed back into the modified equations (12) and (13) from which their magnitudes are evaluated using (9) and (10). Fig. 10 highlights the effect of parasitic loss present. This plot assumes infinite single-ended load impedance, i.e., it represents the input impedances of the balun itself. The effect of loss manifests itself as a degradation of the magnitudes of the even- and odd-mode impedance levels. At resonance, the odd-mode impedance is very high, but finite, and likewise, the opposite is true for the even-mode impedance. These impedances, in parallel with the single-ended load impedance, directly impact the achievable CMRR of the system and also lower the differential power transfer to the next stage. However, the plots of Fig. 10 assumed no capacitors for miniaturization purposes. Normally lumped capacitors will possess finite $Q$ that will further deteriorate the CMRR and reduce power delivery to the next stage. This point is demonstrated by measurements in Section III.

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**Fig. 8.** Calculated magnitudes of the total even- and odd-impedances for a range of single-ended load resistances with circuit parameters $Z_{\text{oe}} = 208 \Omega$, $Z_{\infty} = 51 \Omega$, $Z_{\text{ot}} = 137 \Omega$, and $C = 0 \text{ pF}$.

**Fig. 9.** Calculated magnitudes of the total even and odd impedances for an infinite single-ended load resistance $R_L$ with the following circuit parameters: $\Delta f_1: Z_{\text{oe}} = 157.32 \Omega$, $Z_{\infty} = 56.45 \Omega$, $Z_{\text{ot}} = 90 \Omega$, $C = 0 \text{ pF}$, $\Delta f_2: Z_{\text{oe}} = 20.536 \Omega$, $Z_{\infty} = 47.82 \Omega$, $Z_{\text{ot}} = 100 \Omega$, $C = 0 \text{ pF}$, $\Delta f_3: Z_{\text{oe}} = 30.726 \Omega$, $Z_{\infty} = 22.24 \Omega$, $Z_{\text{ot}} = 80 \Omega$, $C = 0 \text{ pF}$.

for a particular choice of balun circuit parameters. In this example, the magnitudes of the total even- and odd-mode impedances at the crossover frequencies are approximately 43 $\Omega$. However, for a 100-$\Omega$ system, it is obvious (see Fig. 8) that the impedance levels are much lower than 100 $\Omega$ at the crossover frequencies, causing a big mismatch that need to be compensated for by adjusting the element values of the balun. Generally, the higher the system impedance, the higher the coupling required between the coupled lines of the balun.

**E. Effect of Loss on Distributed Biasing**

The remaining and most crucial design parameter to be discussed is the effect of loss of the transmission lines forming the distributed biasing circuit. The circuit $Q$ will be affected by substrate and conductor losses, but mostly by the loss associated
III. EXPERIMENTAL RESULTS

A balanced power amplifier (HELA-10B) from Mini-Circuits,\(^1\) with a gain of 10 dB was selected for experimental proof of principal. The chip is powered by a single +12-V dc power supply via a pair of external lumped inductors and consists of a pair of amplifiers in a pseudodifferential amplifier configuration. As the amplifiers are on the same chip, their gains and phase are well matched. The amplifier has an excellent second-order intercept of 88 dBm and is suitable for broad-band operation from 50 MHz to 1 GHz in a 50-Ω system. A number of distributed biasing circuits were implemented and integrated with the pseudodifferential amplifier. The design objective was to achieve broad-band bandwidth of 1.5:1 high odd-mode and low even-mode RF bias circuit impedances.

A. Conventional Biasing

The amplifier was tested first with the conventional inductor-based biasing circuit. A pair of high-\(Q\) lumped inductors of value 0.75 \(\mu\)H was used and the circuit tested in both the differential and common modes. The measured common- and differential-mode gains are 10 and 9.5 dB, respectively, as shown in Fig. 11(a). The 0.75-\(\mu\)H inductors present very high reactance of approximately 2.3 kΩ, at the center of the band. The inductor values chosen in this example contrast what can be achieved on an RF integrated circuit (RFIC) due to the lower achievable values of spiral inductors and their inherent narrow-band operation due to the self-resonant restraint. However, Fig. 11(a) demonstrates the most vital point about the nondiscriminatory bandwidth produced using the conventional inductor-based biasing scheme.

B. Distributed Biasing Without Miniaturization Capacitors

To verify the properties of the new biasing circuit, the Marchand balun structure with a center frequency of 580 MHz was designed. The optimized balun parameters are \(Z_{\infty} = 208 \Omega\).

\(^1\)Mini-Circuits HELA-10B, 50-MHz–1-GHz amplifier.

on-chip BST capacitor ranges from 30 to 40. Fig. 12(b) illustrates the gains after loading the lumped capacitors with resistors to drop their $Q_s$ to approximately 30. This gives an idea of the effect of loss associated with the capacitors on the system performance. It is obvious from Fig. 12 that the loss degrades the performance of the circuit. In the case of high-$Q$ capacitors, the measured differential-mode gain is 10 dB. However, the common-mode gain is hardly affected and notches at the expected resonant frequency to $-10$ dB. In the case of low-$Q$ capacitors, the common-mode gain only notches to 0 dB and the differential gain reduces from 10 to 9 dB, as expected from the theory. The measured bandwidth in the two cases is 31% of the resonant frequency of 580 MHz, i.e., 1.36 : 1 bandwidth. This demonstrates the effect of loss, as discussed in Section II.

It is also observed from the measured gains of Fig. 12 that the common-mode gain does not notch exactly at the center of the operating band, as was the case without lumped capacitors [see Fig. 11(b)]. This implies that the frequency at which the odd-mode impedance of the balun peaks to a maximum is different from the frequency at which its even-mode impedance is a minimum. This conflict arises due to the addition of lumped capacitors that cause asymmetry in the characteristics of the even- and odd-mode impedances. This may be resolved by fine tuning of the lumped capacitors and optimization of the rest of the circuit elements of the equivalent even- and odd-mode sub-circuits of the balun structure.

![Image](image.png)

**Fig. 12.** Measured common-mode gain $A_c$, and differential-mode gain $A_d$: (a) with a miniaturized balun-like biasing circuit using high-$Q$ lumped capacitors with circuit parameters $Z_{ee} = 208 \, \Omega$, $Z_{oo} = 51 \, \Omega$, $Z_{eo} = 137 \, \Omega$, $C = 5.6 \, \text{pF}$, $R_c = \infty$, and $\tan \delta = 0.016$ and (b) with miniaturized balun-like biasing circuit using high-$Q$ lumped capacitors loaded with resistors with circuit parameters $Z_{ee} = 208 \, \Omega$, $Z_{oo} = 51 \, \Omega$, $Z_{eo} = 137 \, \Omega$, $C = 5.6 \, \text{pF}$, $R_c = 1.2 \, \text{k}\Omega$, and $\tan \delta = 0.016$.

The measured bandwidth, defined as $(\Delta f/ f_0)$, is 34% of the resonant frequency 580 MHz, i.e., 1.4 : 1 bandwidth.

### C. Distributed Biasing With Miniaturization Capacitors

Another design with the same layout dimensions as in (B) was implemented still operating at 580 MHz. The transmission lines were resonant at $3 \times 580 = 1740$ MHz. This yields a reduction in overall size by a factor of three with $C = 5.6 \, \text{pF}$. In Fig. 12(a), the measured response (with high-$Q$ capacitors: $Q \approx 400$) is depicted. It is worth investigating the effect of loading the capacitors with resistors, as this drops their $Q_s$. This was done with the thought in mind that the lumped capacitors could be replaced using tunable components to improve overall system functionality. Thin-film barium strontium titanate (BST) capacitors has recently shown great promise for the construction of high-frequency microwave components [8]. The $Q$ of an

### IV. CONCLUSION

A new distributed biasing circuit has been introduced that replaces conventional biasing using inductors. The physical structure comprises transmission lines and a pair of identical coupled lines in the form of a Marchand balun. The circuit is inherently discriminatory in behavior against the common-mode signals generally originating as substrate noise. This discrimination yields a finite CMRR. However, the differential-mode gain remains, to a great extent, unaffected by the presence of loss in the biasing circuit over a broad frequency range. Possible implementation of this new biasing scheme either on an RFIC or off-chip in the package will greatly improve the overall system performance and eliminates the low-$Q$ narrow-band spiral inductors. The distributed biasing-circuit parameters are fairly straightforward to calculate utilizing expressions based on filter principals to obtain a class of desired impedance functions. Experimental data has successfully verified the theory behind the usage of the balun as a viable biasing circuit.

### REFERENCES


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