Equal capacitor or equal resistor designs can now be easily derived by appropriate replacements in the results for the network of Fig. 4.

V. CONCLUSION

A formula, simpler than the existing one, has been given for the design of the double exponential function \( f(t) = K(e^{-a_1 t} - e^{-a_2 t}) \). Also, the problem of synthesizing an RC two-port for generating \( f(t) \) from an impulse or step excitation has been discussed in detail, and several alternative designs and structures have been presented for this purpose.

REFERENCES


Producing 180° Out-of-Phase Signals From a Sinusoidal Waveform Input

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Abstract—In this paper a simple and novel method for producing two signals with 180° phase difference has been introduced. The phase-balanced signals are produced from a single sinusoidal input signal using two matched op-amps. The phase accuracy of the proposed circuit due to mismatched components is investigated with a two-pole model for the operational amplifiers. With high-precision integrated circuits, the phase error at the amplifier pole frequency (2 MHz) is calculated to be about 0.5° which is a remarkable achievement. The discrete implementation of this scheme is also discussed, and there is a good agreement between the experimental results and the calculated values. The main features of this circuit make it suitable for use in bridge measuring systems and lock-in amplifiers.

I. INTRODUCTION

In many instrumentation systems, the existence of two 180° out-of-phase sinusoidal signals is essential. These two signals could be used for increasing the sensitivity of bridge circuits such as a Wheatstone bridge, which can be used for measuring an unknown resistor in strain gauge systems [1]. The other great advantage of two well-balanced 180° out-of-phase signals is the ability to eliminate the effects of offset in lock-in-based systems [2]. The accuracy and reliability of such signals are, therefore, very important in the performance of high-precision measuring instruments. At low frequencies, generating such signals is not a complicated task, but for high frequencies the effects of nonideal circuit elements will cause deterioration from the ideal response. Two traditional remedies exist for this problem; the first is to use a transistor pair as a differential amplifier, and the second is to use an operational amplifier in the inverting mode. The former is a well-known method, but it has several limitations such as low permissible input voltage levels, and the requirement of finding two highly matched transistors. The main problem with operational amplifiers is low gain-bandwidth product (GBWP). In this paper we model the amplifier performance and minimize phase errors by using matched op-amps.

II. THEORETICAL ANALYSIS

To develop the theory for our two-matched-op-amp model, we start with a simple first-order transfer function for a conventional op-amp-based inverter for producing two 180° out-of-phase signals (Fig. 1(a)). This method has, however, major limitations at high frequencies. We can write down the overall input-output characteristic of such a simple inverter as

\[
\frac{V_o}{V_{in}} = -R_2 \frac{B}{R_1} \left( \frac{1}{R_2/(R_2 + 1) + s} + \frac{B}{s} \right)
\]

where \( B \) is the GBWP of the op-amp. As it is apparent in (1), the transfer function of the inverter has a pole at \( \omega = B/(R_2/(R_2 + 1)) \). This pole not only causes a decrease in the gain at the frequencies above it, but also degrades the output phase characteristic at frequencies much below it. To reduce this effect several methods have been tried.

One of the most important methods is to use active feedback with matched operational amplifiers [3], and another approach is to use composite operational amplifiers [4]. These methods are useful in situations where we need to produce a 180° out-of-phase signal from a signal that must maintain its phase integrity and can not be manipulated by any means. However, some situations exist where these two signals are fed into the front-end of the system, so the initial phase of the feeding signal is not important, and the only important factor is the phase difference between these two signals. For this purpose, the use of two matched op-amps (where one of them inverts the input signal, with a residual phase error caused by the nonideal op-amp characteristics, and the other produces a signal with a similar residual phase error) seems to be useful.

Fig. 1(b) shows a simple proposed circuit for producing two 180° out-of-phase signals. As mentioned before, the phase difference between \( v_{o1} \) and \( v_{o2} \) is just equal to 180°, if \( a_1 = 0 \) and the two op-amps match exactly. The two-pole transfer function for each op-amp can be written as

\[
G(s) = \frac{B_\alpha}{s(s + \alpha)}
\]

where \( \alpha \) is the second pole. We can then write down the transfer functions of these two amplifiers as

\[
H_1(s) = \frac{a_1 + 1}{s(a + a_1)} + \frac{B_\alpha}{s} = \frac{B_\alpha}{s(a + a_1)} + (a_1 + 1)
\]
Fig. 2. Deviation of the output phase difference from 180° for the two-pole approximation and \(\Delta B/B = 2\% - 10\%\) with a step of 2% (from bottom to top, respectively).

\[
H_2(\omega) = \frac{(-a_2)}{(\omega^2 + a_2^2)}.
\]

Hence the phase deviation from 180° can be written as

\[
\Delta \Phi_1 = \Delta \Phi - \pi = \text{arg}[H_2(j\omega)] - \text{arg}[H_1(j\omega)]
\]

where \(\Delta \Phi\) is the phase difference between the two outputs.

Now we want to take into account the effects of mismatch in the three pairs \((a_1, a_2), (B_1, B_2),\) and \((P_1, P_2)\). For this purpose we use the relationships \[5\]

\[
\begin{align*}
(a_1 + a_2)/2 &= a, \quad a_1 - a_2 = \Delta a \\
(B_1 + B_2)/2 &= B, \quad B_1 - B_2 = \Delta B \\
(\omega_1 + \omega_2)/2 &= \omega, \quad \omega_1 - \omega_2 = \Delta \omega
\end{align*}
\]

By substituting (3), (4), and (6) into\(5\) we obtain

\[
\Delta \Phi_1 = \frac{P_0B}{[BP - (a + 1)\omega^2]^2 + (a + 1)^2P^2\omega^2}.
\]

In our analysis we have equated \(\Delta \Phi_1\) by its tangent because \(\Delta \Phi_1 \ll 1\). Since \(P\) and \(B\) belong to op-amps with a similar manufacturing process we can assume \(\Delta P/P = \Delta B/B\). Moreover, for the case of \(B = P\) (close approximation), we can rewrite \(\Delta \Phi_1\) for the worst case of mismatch between two op-amps as

\[
|\Delta \Phi_1| = \frac{[\Delta a] + (\omega_1^2\omega_2^2 + (\omega_1^2 + \omega_2^2)]}{[B - (a + 1)/\omega^2] + (a + 1)^2\omega^2}.
\]

For the case of \(\omega \ll B\), (8) reduces to a similar relation similar to that of the one-pole model.

### III. RESULTS

Achieving values of \(|\Delta a|/a = 0.1\%\) is not unreasonable; therefore \(|\Delta B|/B\) (the mismatch of the two op-amps GBWP) is the ultimate source of error. Fig. 2 shows the variation of \(\Delta \Phi_1\) vs. frequency for several values of \(\Delta B/B\) for two typical op-amps (such as the LF347) with \(B = 8\pi \times 10^6\) rad/s. The resistors have been chosen for \(a = 1\) and \(\Delta a/a = 0.5\%\). The maximum value of \(\Delta \Phi_1\) occurs approximately at the amplifier pole frequency \(B/(a + 1)\), which is 2 MHz in this example. Also for highly matched op-amps \((\Delta B/B = 2\%\) and resistors \((\Delta a/a = 0.1\%\), the \(\Delta \Phi_1\) does not exceed 0.5 degree. At the frequencies greater than the amplifier pole-frequency, the gain of these two amplifiers decreases; therefore the reliable range of operation extends only up to the amplifier pole-frequency.

One of the most important parameters is the ratio factor of the resistors, \(a\). The importance of this parameter is obvious from (8), so by decreasing its value we can decrease \(\Delta \Phi_1\). Fig. 3 shows \(|\Delta \Phi_1|\) for \(a = 0.3, 0.65, 1\) with \(\Delta B/B = 2\%\) and \(\Delta a/a = 0.5\%\). Decreasing \(a\) also increases the operating range of the circuit.

In order to test the proposed circuit, the discrete version of the circuit was assembled with a popular op-amp (LF347) and two simple 3.3 k\(\Omega\) resistor arrays (with about 0.5% mismatch). A function generator (Iwatsu FG-330) was used to provide an input sinewave signal for the circuit at a frequency range of 1 kHz to 10 MHz. The outputs of the circuit then were connected to a fast oscilloscope (Iwatsu SS-5421, 350 MHz) for measurements. By changing the input signal frequency the delay time between the two output signals was measured accordingly.

The phase deviation from 180° between the two outputs has been obtained by multiplying this delay time by the signal frequency, \(\omega\). In order to reduce the error at low frequencies (small phase deviations), the average value of several measurements was considered for the phase calculation. By this method a minimum phase deviation of about 0.1° can be achieved. The phase deviations from 180° as a function of frequency were obtained, and the results are shown in Fig. 4. For a better comparison, the estimated theoretical values and the phase response of a typical conventional op-amp-based inverter are also shown in the same figure.
As can be seen from Fig. 4, the general behavior of the experimental curve is very similar to that obtained theoretically, but the phase deviation values are slightly higher than the calculated values. The difference may be due to the higher degree of mismatch in the resistors and op-amps, stray capacitances and the frequency behavior of the resistor arrays. However, the phase deviation is substantially improved over the conventional circuit. With values of about 10% for $\Delta B/B$, and 0.5% for $\Delta a/a$ (which are reasonable values for commercial components), there is a good agreement between the theoretical and the experimental results as can be seen in Fig. 4.

IV. CONCLUSION

A simple and accurate method for producing 180° out-of-phase signals from a single sinusoidal input has been described. Simple phase formulas for the two-pole scheme have been derived, and the role of mismatch parameters has been investigated. The main results of this study are: 1) The proposed scheme shows a significant advantage over the conventional op-amp-based inverter. 2) The effects of mismatch parameters, $\Delta B/B$, and $\Delta a/a$, on the output phase error indicate that both $\Delta B/B$ and $\Delta a/a$ have a major role in the phase errors, but $\Delta a/a$ can be controlled. 3) From Fig. 4 it is obvious that in spite of the improvement over a conventional op-amp-based inverter, the implementation of the reported circuit using commercial components is suitable for frequencies only up to 400 kHz. Therefore for a phase error of less than 1°, circuit operation should be restricted to one-twentieth of the GBWP of the commercial op-amps.

REFERENCES


High Current Injection Effects of the Switching JFET's in the Current Electrometers

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Abstract—The performance of JFET as a switch in the feedback loop of a multigain ranging current electrometer is described. The study indicates that the electrometer starts departing from linear behavior as the input current becomes comparable to the drain saturation current of the JFET. It is shown that the nonlinearity in the electrometer behavior is attributable to the variations occurring in the ON resistance and gate current of the JFET.

Fig. 1. Basic electrometer configuration.

I. INTRODUCTION

A current electrometer is a current-to-voltage converter, commonly used to measure very low currents, typically in the $\mu A$ range and below. Electrometers find applications in a variety of measuring instruments in many diverse areas such as mass spectrometry, particle accelerators, ultra-high-vacuum technology, semiconductor measurements, photometric measurements, atmospheric research etc. [1]–[4]. In a majority of these applications the current to be measured at any instant of time may vary by as much as 5 to 6 decades. This can be accomplished by using either a logarithmic electrometer or a linear electrometer. In order to measure both the low and high currents accurately over a wide dynamic range of signal current with linear electrometers, it is necessary to provide ranging to the electrometer. Ranging can be accomplished by suitably changing the feedback in the electrometer amplifier, shown in Fig. 1. In this figure $R_f$ is the feedback resistance, and the JFET functions as a switch for gain ranging operation.

The dynamic range of useful operation of the electrometer in our configuration is primarily decided by the switch. In this article, we analyze the behavior of the JFET for its influence on the electrometer performance. We chose the 2N 4117 A for its low leakage current that allows measurements down to 1 pA. We present both theory and practical results.

II. JFET Switch Properties

Refer to Fig. 1. The gate of the JFET is pulled up by a resistance $R_g$ for providing suitable bias for the ON-condition. The source terminal is held at virtual ground potential through the operational amplifier.

The input current flows through the JFET and causes a voltage $V_d$ to appear at the drain terminal. It will be shown that $V_d$ is the main source of nonlinearity in an otherwise linear relationship between input current and electrometer output. The magnitude of nonlinearity depends on the magnitude of input current and JFET ON resistance.

In the following sections, using the JFET theory, this relationship will be examined. Two separate cases are considered viz., the majority carrier injection effects and the minority carrier injection effects.

A. Majority Carrier Current Injection

The flow of majority carrier current $I_m$ through the JFET channel results in a positive $V_d$ such that reverse biasing of the gate-drain junction occurs. In that case, $I_m$ equals device leakage current (<1 pA). When $V_d$ is less than device pinch-off voltage $V_p$, JFET