ES 272 Assignment #4

Due: 10:00am, March 19th, 2009

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(Problem 1*) Narrowband RF tuned-tank amplifier design (100pt)

Figure 1 shows a single-ended tuned-tank MOS amplifier including a bias circuit that consists of M3, R_1, R_2, and C_g. Using the $nmos2v$ transistors in the $tsmc18rf$ library, design the circuit at around 5 GHz. The voltage gain (from the input terminal to the output terminal in the figure) should be *at least* 30 and the input impedance shown in the figure should be 50 Ω with negligible reactance at the design frequency (5 GHz). The maximum Q that you can use for the LRC tank is 10. For any inductor in the circuit, maximum inductance value that you can use is 5 nH. Use V_{dd}=1.8 V. The gate oxide thickness, t_{ox}, is about 4.1 nm for the MOS transistors in 0.18 μm technology, and this information may be needed in at least roughly calculating the gate-source capacitance in pinch-off, where the formula is $C_{gs} = (2/3) \cdot LWC_{ox}$ ignoring the overlap capacitance between the gate and the source. Here, $C_{ox} \approx 3.9 \cdot \epsilon_0 / t_{ox}$.

Evaluate the linearity of your designed circuit by simulating the input 1-dB compression point at 5 GHz, and also by simulating the IIP3 point using two harmonic tones of your choice around 5 GHz.

Estimate the noise figure of your designed amplifier, only considering 3 noise sources, the termanl drain noise (use $\gamma = 3$) of transistors M1 and M2 and the Johnson-Nyquist noie of the tank resistor R.

What is the spurious free dynamic range (SFDR) of your amplifier?

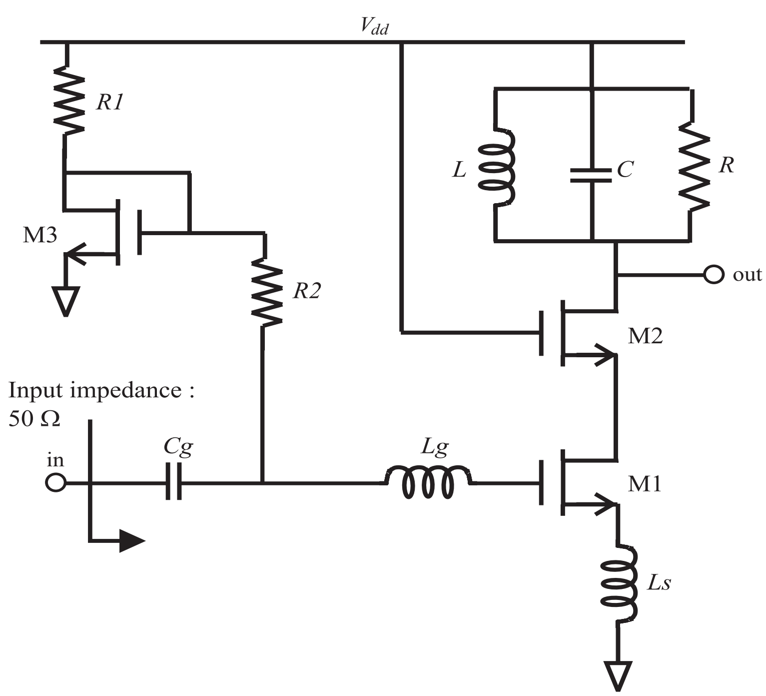

Figure 1: Problem 1

(Problem 2) Quadrature generation using RC polyphase filters (30pt)

The generation of quadrature signals (sine and cosine with the same frequency) is essential to perform vector modulation and demodulation in wireless transceivers. We have seen the use and usefulness of quadrature signals in the context of receivers (Lecture note #1, page 2). We will talk more about quadrature signals as we go along in this class. This problem is concerned with a quadrature generation technique using RC filters.

(a) *RC-CR phase splitter:* One way to generate quadrature signals is to use an RC-CR phase-shift network shown in Fig. 2(a). V is an oscillator input signal, from which output signals, V_1 and V_2, are generated. Calculate V_2/V_1 in the frequency domain, and show that V_1 and V_2 have a phase difference of 90° at all input frequencies, but the amplitudes of V_1 and V_2 have the same magnitude only when the frequency of V is given by $\omega = 1/(RC)$.

(b) *1st-order RC polyphase filter:* Figure 2(b) shows a 1st-order RC polyphase filter. Two input terminals are driven by differential oscillator signals (V and $-V$) whose frequency is ω while the other two input terminals are tied to ground. Show, by calculating V_2/V_1 in the frequency domain, that V_1 and V_2 have a phase difference of 90° at all input frequencies, but their amplitudes have the same magnitude only when $\omega = 1/(RC)$. In this 1st-order polyphase filter, as the frequency deviates from $1/(RC)$, the amplitude mismatch between the quadrature signals increases relatively rapidly. You can mitigate this problem and make a broader band quadrature generator by increasing the order of the polyphase filter. See the next problem.

(c) *2nd-order RC polyphase filter:* Figure 2(c) shows a 2nd-order RC polyphase filter. Again among the four input terminals, two terminals are driven by differential oscillator signals (V and $-V$) whose frequency is ω while the remaining two terminals are connected to ground. Demonstrate, by calculating V_2/V_1 in the frequency domain, that V_1 and V_2 have a phase difference of 90° at all input frequencies, but their amplitudes have the same magnitude only when $\omega = 1/(RC)$. (*Hint:* Note that the input signals at the four input terminals, $(V, 0, -V, 0)$, can be decomposed into $(1/2) \times (V, Ve^{j\pi/2}, Ve^{j2\cdot\pi/2}, Ve^{j3\cdot\pi/2})$ and $(1/2) \times (V, Ve^{-j\pi/2}, Ve^{-j2\cdot\pi/2}, Ve^{-j3\cdot\pi/2})$. In either of these two vectors, the relative phase difference between any neighboring components is $\pm\pi/2$. You may exploit this constant phase difference and the physical symmetry of the polyphase filter to simplify your calculation.) Can you see that as compared to the 1st-order filter, the amplitude mismatch between V_1 and V_2 has less dependence on the frequency offset from $\omega = 1/(RC)$ in the 2nd-order polyphase filter? Best way to show this is to plot $|V2/V1|$ vs. ω around $\omega = 1/(RC)$ for both the 1st- and 2nd-order polyphase filter.

(d) *3rd-order RC polyphase filter:* You can repeat the procedure in (c) and show that the ratio of the quadrature signals (V_1 and V_2) in the 3rd-order polyphase filter shown in Fig. 2(d) is given by (You don't have to derive this, but if you are interested, you can systematically tackle this problem using a difference matrix equation.)

$$\frac{V_2}{V_1} = j \cdot \frac{3\omega RC + (\omega RC)^3}{1 + 3(\omega RC)^2} \tag{1}$$

Check again that the two signals have 90 ° phase shift at all frequencies, and their magnitudes are the same only when $\omega = 1/(RC)$. Plot $|V_2/V_1|$ vs. ω for the 3rd-order polyphase filter and compare it to the plots of the previous problem (c). You should be able to see that the 3rd-order polyphase filter can generate quadrature signals over a broader band than the 2nd-order polyphase filter.

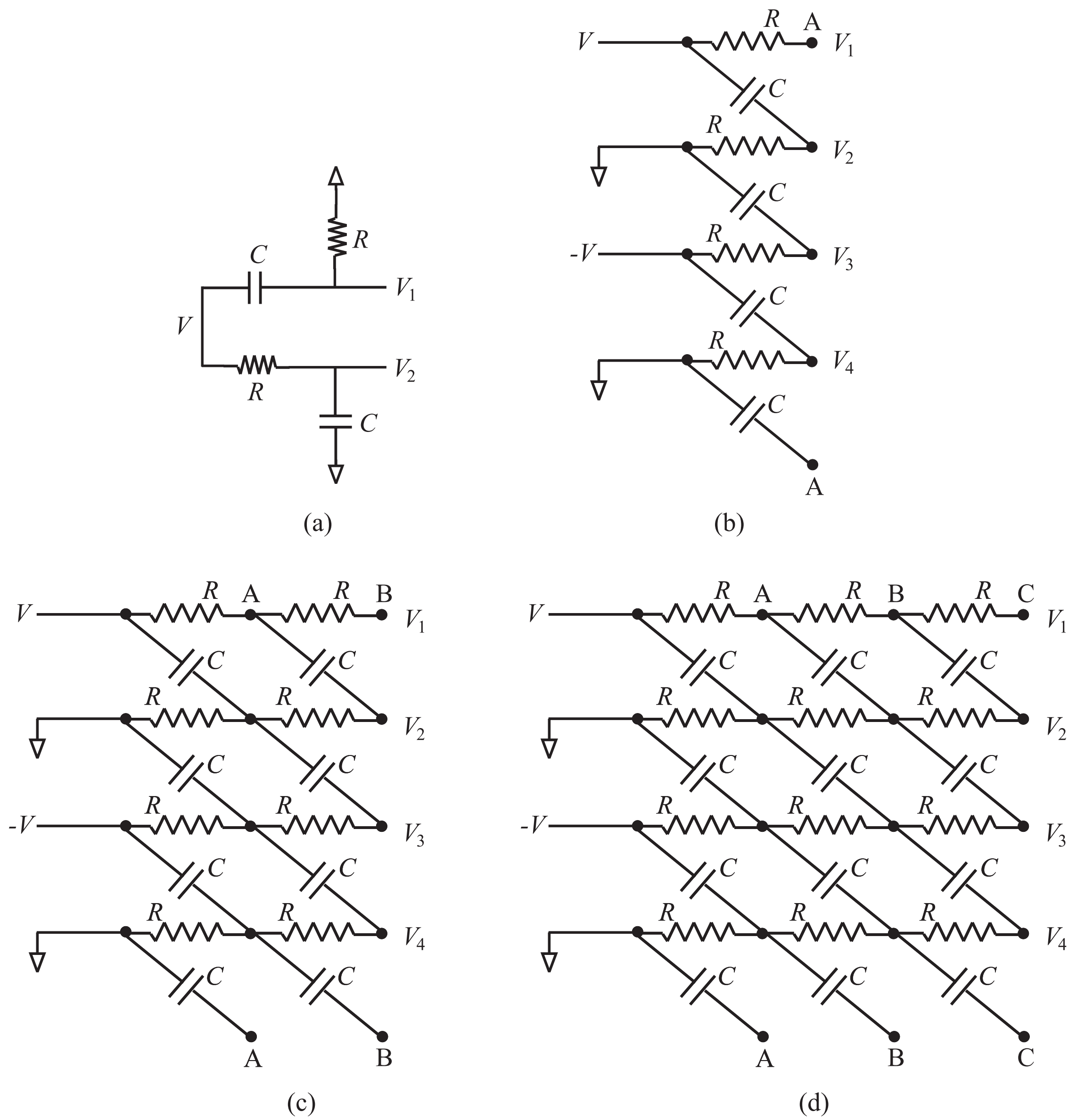

Figure 2: Problem 2: (a) RC-CR network. (b) 1st-order RC polyphase network. (c) 2nd-order RC polyphase network. (d) 3rd-order RC polyphase network.