VGA DC-offset calibration and DC Servo Loop Design

- **DC offset in VGA/PGA**
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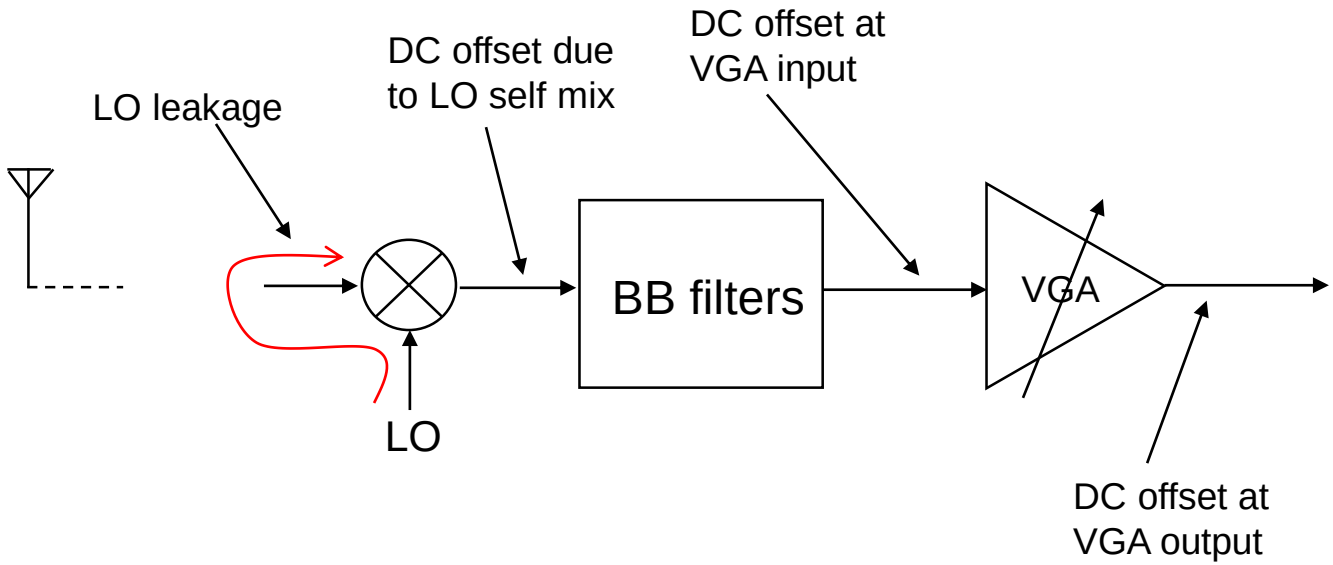
- **References**

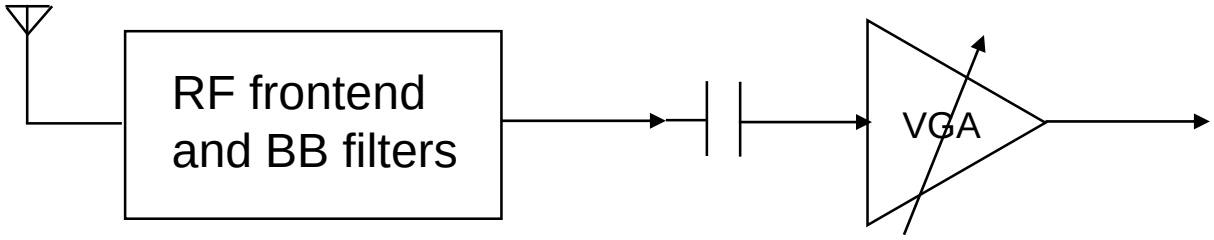
The problem of DC offset in baseband VGAs



- DC offset at VGA input can be due to:
  - LO self mixing at the RF mixer with DC propagating all the way to VGA
  - DC offset due to mismatch in the baseband filter and any preceding baseband circuits (mixer output for example)
  - DC offset due to VGA circuit itself
- for 60dB max gain VGA, the DC offset at the VGA output is 1V for only 1mV DC offset at the input!

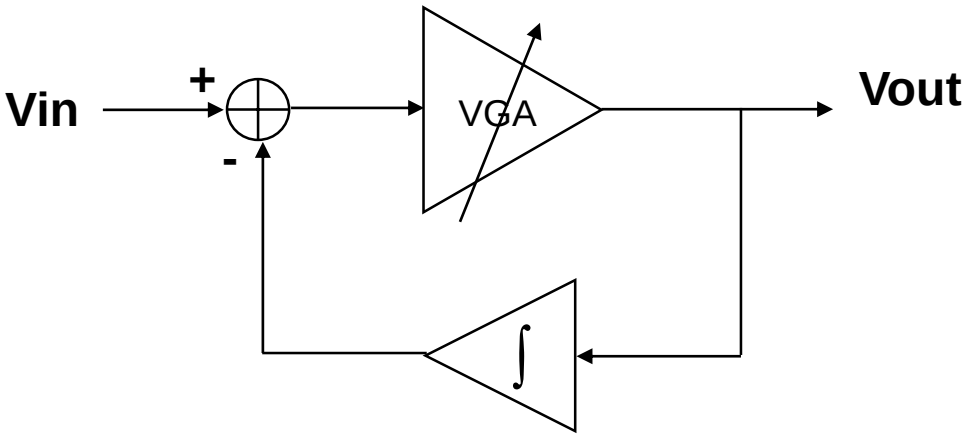
How to get rid of DC offset?

1. AC coupling:



- AC coupling removes DC offset due to preceding stages, but does nothing to DC offset generated within the subsequent stages. For example, AC coupling the BB filter to VGA blocks the filter offset but does not impact offset due to VGA itself
- The highpass corner frequency is set by the coupling cap and the VGA input impedance, which could be temperature and process dependent. A better technique is to force the highpass corner to depend on an on-chip resistor with VGA input impedance set to be very high (10x).

2. Use servo loop:



- a servo loop is just an integrator put in the feedback loop between the VGA output and its input as shown. The new input-output relation can be derived as

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA}(s)}{1+T}$$

$$T = A\frac{G_{VGA}(s)}{s} \quad ; \text{where A/s is the integrator transfer function}$$

$$\Rightarrow \frac{V_{out}(s)}{V_{in}(s)} = \frac{sG_{VGA}(s)}{s + AG_{VGA}(s)}$$

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Note that at DC, $s \rightarrow 0$, $V_{out}/V_{in} = 0$, which what we want. At frequencies much higher than DC, $s \rightarrow \infty$, $V_{out}/V_{in} = G_{VGA}$, which is the VGA forward gain with the servo circuit completely out of the picture.

Problem!

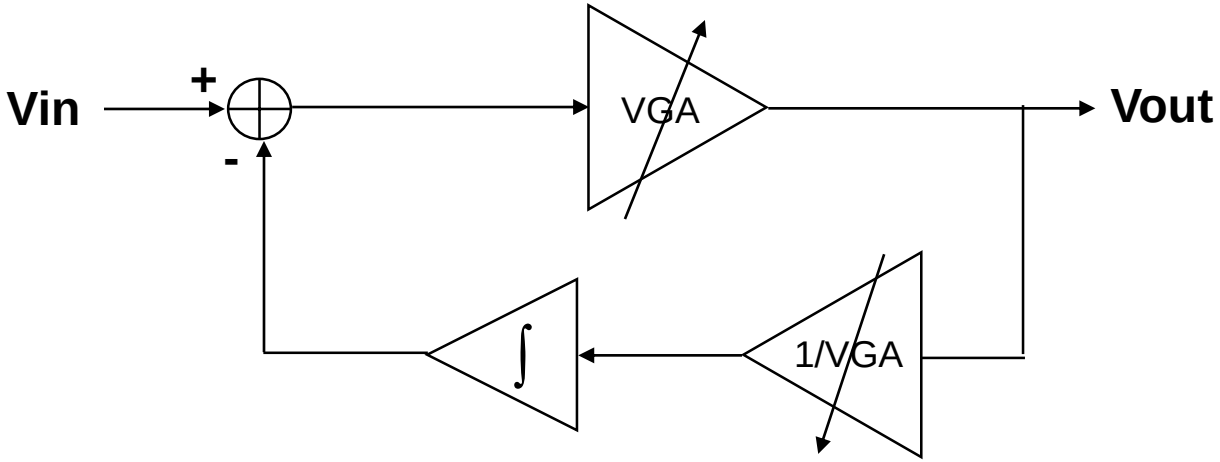
In the above equation, the highpass pole is given by

$$s + AG_{VGA}(s) = 0$$
$$\Rightarrow P_{HP} = -AG_{VGA}$$

The pole is gain dependent! This means, whenever the VGA changes gain, the highpass pole simply shifts! Not good.

→ design gain-independent highpass servo loop

Gain-independent highpass corner servo loop:



The idea is to insert a block in the feedback loop whose function is the inverse of the VGA forward gain. In other words, this block has the inverse AGC slope of the forward path VGA. The new transfer function is then:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA}(s)}{1+T}$$

$$T = A\frac{G_{VGA}(s)}{s}\frac{\mathrm{K}}{G_{VGA}(s)} \quad ; \text{where A/s is the integrator transfer function}$$

$$\Rightarrow \frac{V_{out}(s)}{V_{in}(s)} = \frac{sG_{VGA}(s)}{s+AK}$$

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Note that at DC, s→0, V_{out}/V_{in} = 0, just like before. At frequencies much higher than DC, s→∞, $V_{out}/V_{in} = G_{VGA}$, which is the VGA forward gain with the servo circuit completely out of the picture. Now, the highpass pole is:

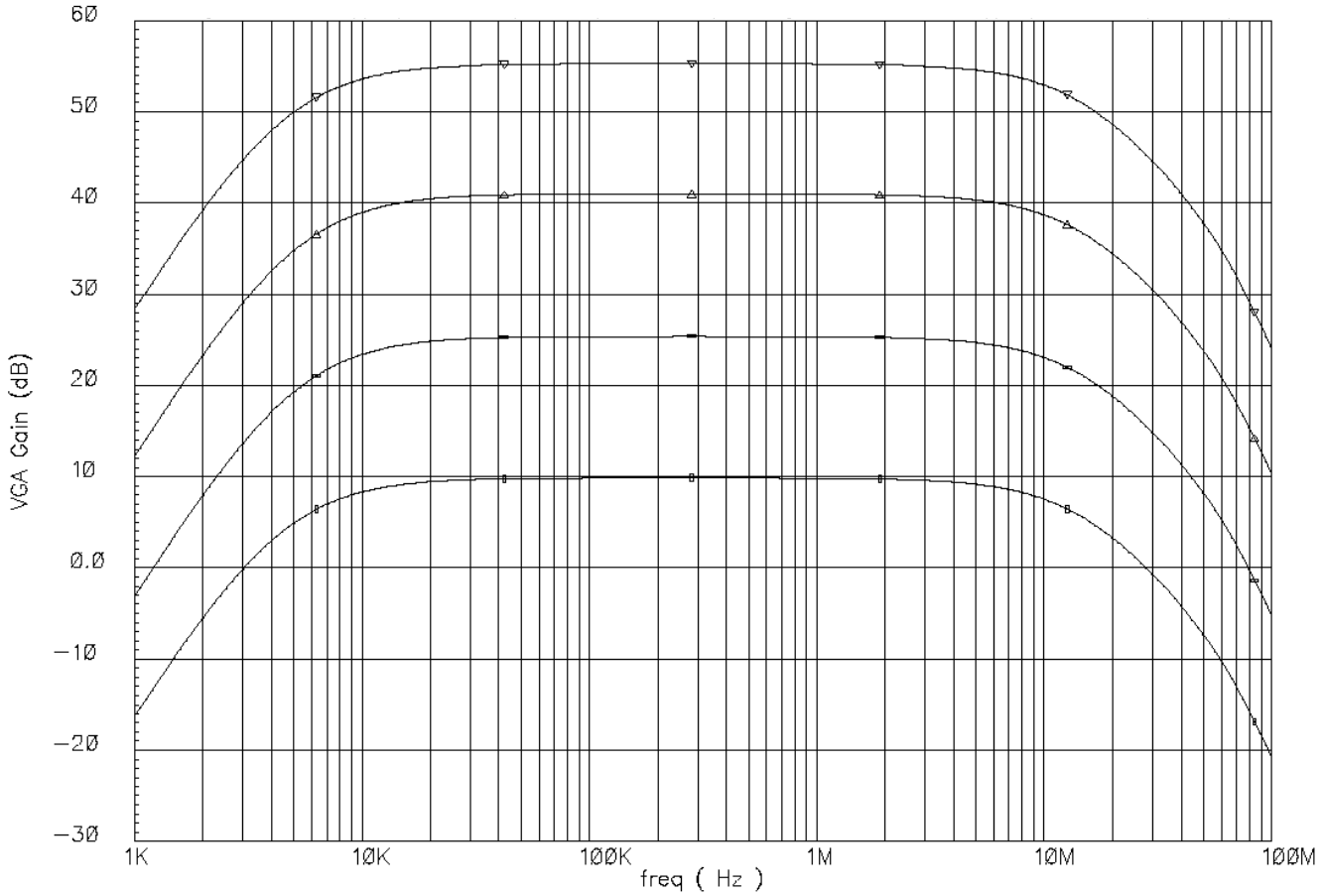
$$s + AK = 0$$

$$\Rightarrow P_{HP} = -AK$$

The highpass pole is fixed and is gain independent.

Note that the inverse VGA block does not have to be an exact replica of the forward VGA block. It only needs to have the inverse AGC slope.

Simulated VGA gain over freq and over AGC:



- With modified servo loop scheme, <10% variation of high-pass corner over 60dB AGC is obtained

Impact of nonideal integrator:

In the previous equations we assumed an ideal integrator in the servo loop with infinite DC gain. The impact of finite gain integrator on the servo behavior.

Let us assume the nonideal integrator has the following function:

$$H_{\text{int}}(s) = \frac{A}{s + P_0}$$

Therefore, the integrator DC gain = A/P_0. Substituting in the VGA/servo equation we get:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA}(s)}{1+T}$$

$$T = A\frac{G_{VGA}(s)}{s + P_0}\frac{\text{K}}{G_{VGA}(s)}$$

$$\Rightarrow \frac{V_{out}(s)}{V_{in}(s)} = \frac{(s + P_0)G_{VGA}(s)}{s + (P_0 + AK)}$$

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{(s+P_0)G_{VGA}(s)}{s+(P_0+AK)}$$

Note that at DC, s→0

$$\frac{V_{out}(DC)}{V_{in}(DC)} = \frac{P_0 G_{VGA0}}{(P_0+AK)}$$

It is seen that the nonideal integrator creates some residual DC offset at the VGA output. The residual DC offset gets worse with increased AGC and improves with higher integrator DC gain, *A*.

At frequencies much higher than DC, s→∞, $V_{out}/V_{in} = G_{VGA}$, which is the VGA forward gain with the servo circuit completely out of the picture, which means the nonideal integrator has no impact there.

The nonideal integrator also slightly shifts the highpass pole a little by its own pole, P_0. This is not an issue as long as $P_0 << AK$.

Stability:

As in any feedback system, the VGA/servo feedback system is of stability concern. The ideal transfer function of this arrangement is seen to be

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA}(s)}{s + AK}$$

Which is unconditionally stable, assuming the VGA itself is stable for all frequencies. However, with the VGA being band limited and the integrator is not ideal, a more realistic transfer function of the VGA/servo loop would be:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA}(s)}{1+T}$$

$$G_{VGA}(s) = \frac{G_{VGA0}}{s + P_{VGA}}$$

$$\frac{K}{G_{VGA}(s)} = \frac{K}{G_{VGA0}(s + P_{invVGA})}$$

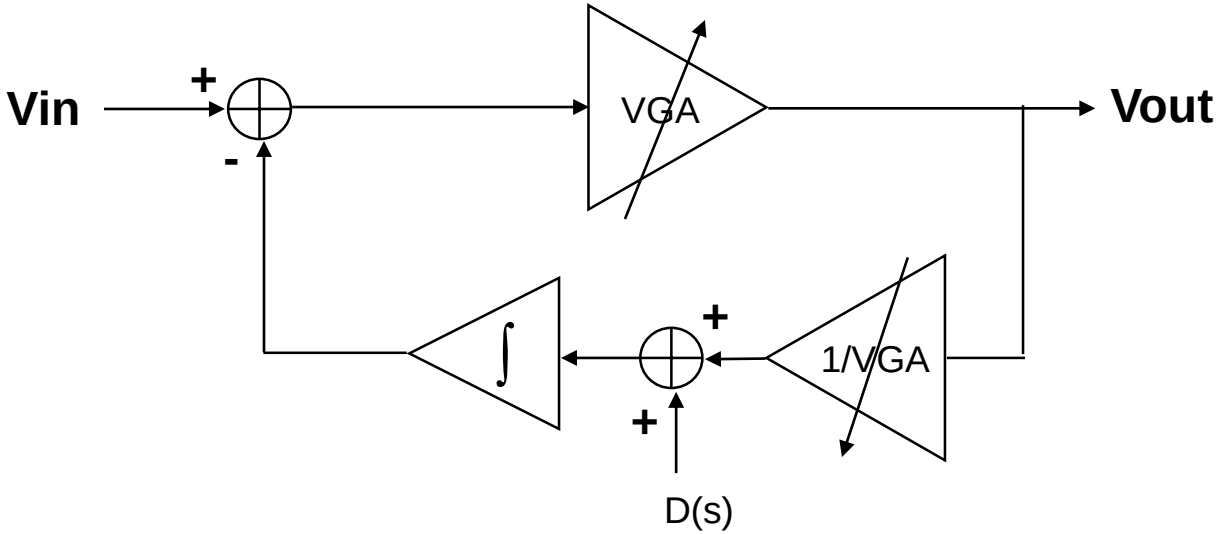
$$T = A\frac{G_{VGA}(s)}{s + P_0}\,\frac{\mathrm{K}}{(s + P_{VGA})(s + P_{invVGA})G_{VGA}(s)}$$

$$\Rightarrow \frac{V_{out}(s)}{V_{in}(s)} = \frac{G_{VGA0}\left(s^2 + (P_0 + P_{invVGA})s + P_0 P_{invVGA}\right)}{s^3 + \left(P_0 + P_{invVGA} + P_{VGA}\right)s^2 + \left(P_{invVGA}P_{VGA} + P_{VGA}P_0 + P_0 P_{invVGA}\right)s + \left(P_{invVGA}P_{VGA}P_0 + AK\right)}$$

Here we assumed both the forward path VGA and the inverse feedback block are of a single pole approximation. In reality, this might not be true, especially for multi-stage VGAs. In this case both VGAs will have more complex transfer function with multiple poles that could make stability of more concern.

As seen, a more realistic transfer functions shows a third order function which is prone to oscillation if not carefully designed.

Distortion analysis of servo loop:



Let us the investigate the impact of the 1/VGA nonlinearity on the overall VGA distortion. The distortion due to the 1/VGA block can be injected into the loop as shown above. This distortion then reaches the VGA output through the following transfer function:

$$\frac{V_{out}(s)}{D(s)} = \frac{\frac{-AG_{VGA}(s)}{s}}{1+T}$$

$$T = A\frac{G_{VGA}(s)}{s}\frac{\mathrm{K}}{G_{VGA}(s)} \quad ; \text{where A/s is the integrator transfer function}$$

$$\Rightarrow \frac{V_{out}(s)}{D(s)} = \frac{-AG_{VGA}(s)}{s+AK}$$

The distortion reaches the output via a lowpass function whose corner frequency exactly equals the highpass pole of the entire VGA/servo arrangement. This means the 1/VGA distortion is greatest around the edge of the highpass corner, and progressively gets attenuated as the frequency gets higher.

Note that in our analysis, we assumed the distortion to be small enough so that the s-domain representation is still valid. In this case, distortion is treated as “noise”, similar to quantization distortion in A/D. Although this is not a very good assumption, never the less the analysis gives an insight on what is going on. If the 1/VGA block distortion is significantly high, the above analysis might not be accurate and should be taken with a grain of salt.

Impact of servo loop noise on overall system:

Similar to distortion analysis, the 1/VGA noise reaches the output through the same lowpass function:

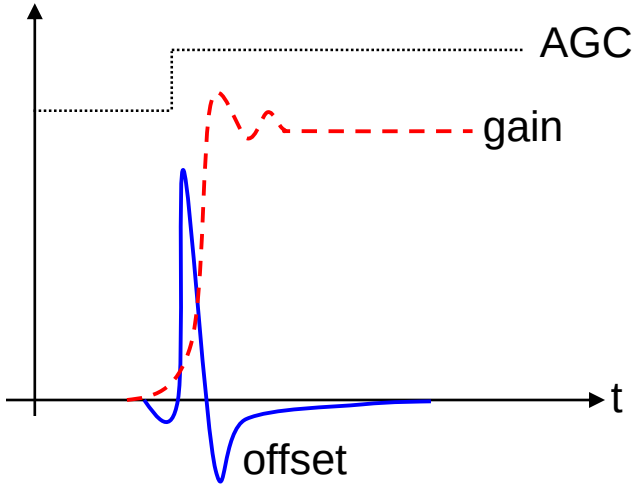
$$\frac{V_{out}(s)}{n_{invVGA}(s)} = \frac{-AG_{VGA}(s)}{s+AK}$$

Therefore, for frequencies higher than the highpass corner, the noise from the 1/VGA block diminishes. The noise from the integrator, however, in not distinguished from the VGA input, and so passes to the output through the highpass function:

$$\frac{V_{out}(s)}{n_{\text{int}}(s)} = \frac{sAG_{VGA}(s)}{s+AK}$$

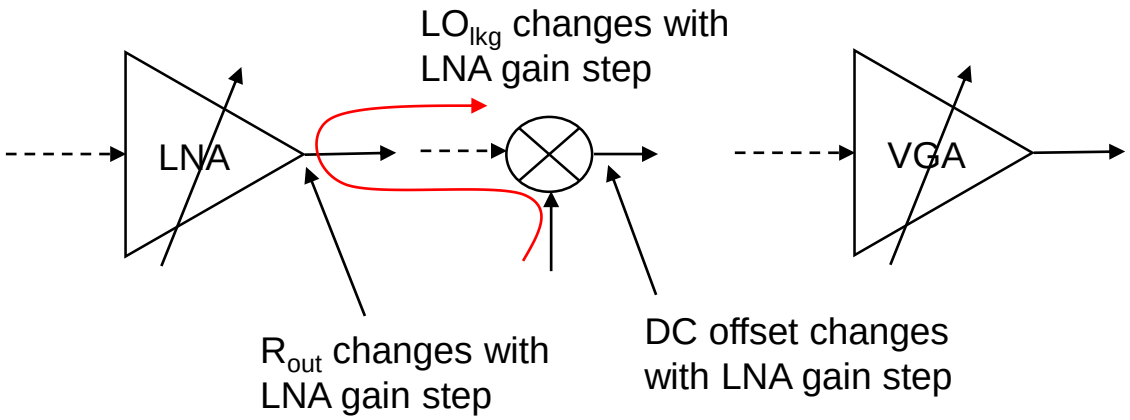
However, it is important to note that the integrator noise itself is shaped due to the integration function. In other words, the integrator output noise has an 1/f shape. Therefore, with the highpass function of the VGA/servo loop, the integrator noise is less of a concern.

The choice of highpass corner:



any sudden change in the AGC voltage due to response of AGC control loop to a fading condition for example, results in a DC offset glitch, which settles down by a time constant equals to the RC time-constant of the highpass corner. The higher the highpass corner frequency is the faster this glitch settles down. However, some of the demodulated signal energy resides close to DC, and so there is always a limit on how high a highpass corner can get before it starts to attenuate part of the desired signal itself at basband. Moreover, the highpass corner creates group delay ripple at the lower edge of the signal band that would degrade EVM if the corner is too close to the band edge

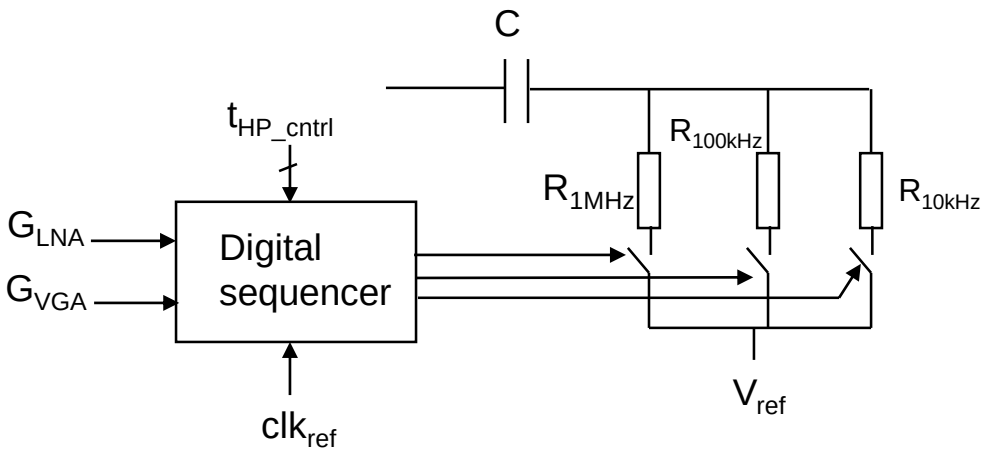
Events causing DC offset glitch in ZIF receiver:



● the LNA output impedance changes slightly with LNA gain switch. This modifies a bit the reflected component of the LO leakage, resulting in a small change of DC offset at the VGA input. This results in a nasty DC offset kick at the VGA output, especially at max gain, before the servo loop has the time to correct for it.

● a sudden change in the VGA gain over 20dB step or so (due to signal search by DSP or due to signal fading), results also in a DC offset kick, especially in AGC dependent DC offset VGA topology (such as current steering). In many cases, VGA/PGA input DC offset is PGA/VGA gain-dependent

Techniques to speed up DC offset settling:



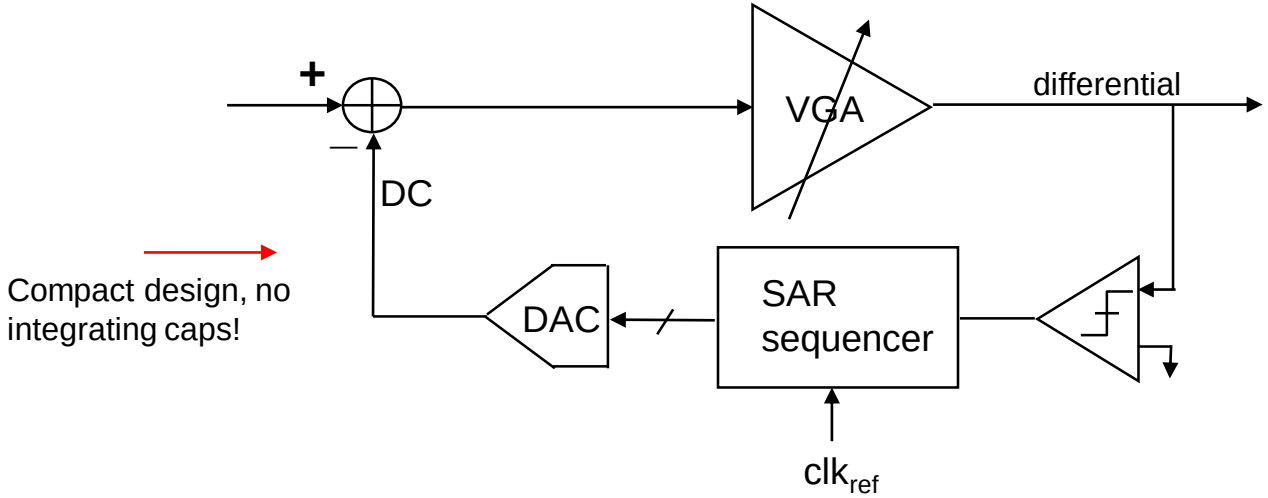
● the use of a switchable highpass corner controlled by a digital sequencer each time an “event” of LNA gain switch of VGA switch occurs.

● in the example scheme above, first, the 1MHz corner is switched on for one or two time constants for very fast coarse DC offset settling. Then the 100kHz corner is switched on for also one or two time constants for fine offset settling before the final 10kHz corner is switched on permanently until the next event occurs.

● the time during which each of the switcheable highpass corners lasts is made programmable by the DSP firmware.

● very fast settling can be achieved using this technique.

The use of digital servo loop: most popular method



● a comparator detects the differential output offset and its sign. Based on the offset sign, the comparator triggers a successive approximation register sequencer to start counting up or down

● a DAC converts the SAR digital output into DC offset current/voltage dumped at the VGA input (usually opamp virtual ground terminals) to null out offset.

● the SAR takes n/2+1 clks to reach decision until the output differential offset changes sign and the comparator output changes. The final value is then latched and stored in memory.

● the cycle is repeated for every gain setting. This method has the advantage of no large RC time constant involved nor any transients. However, loop does not track offset drift due to temperature for example.

● you need to watch for the DAC noise as it could impact VGA noise if not designed carefully.

References:

[1] C. Hull, R. Chu, J. Tham, “A direct-conversion receiver for 900MHz(ISM band) spread-spectrum digital cordless telephone”, IEEE JSSC, vol. 31, pp. 1955 - 1963, December 1996.