

*Getting started with the
gnt analysis in Cadence Virtuoso*
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The latest version of this document can be found on www.analogdesign.be

Need help?

This document assumes basic knowledge about Cadence Virtuoso, the schematic editor, the simulation environment Analog Design Environment (ADE) and the visualization tool ViVA. More information can be found in the help files included in the Virtuoso installation.

For information about the theory behind the general network theorem (GNT), general feedback theorem (GFT), extra element theorem (EET) and chain theorem (CT) and their implementation in Virtuoso, go to analogdesign.be, where you'll also find installation instructions and further documentation.

Found an issue?

If you encounter a bug, please add an issue to the public issue tracker at

<https://bitbucket.org/analogdesign/gnt/issues?status=new&status=open>

Work in progress

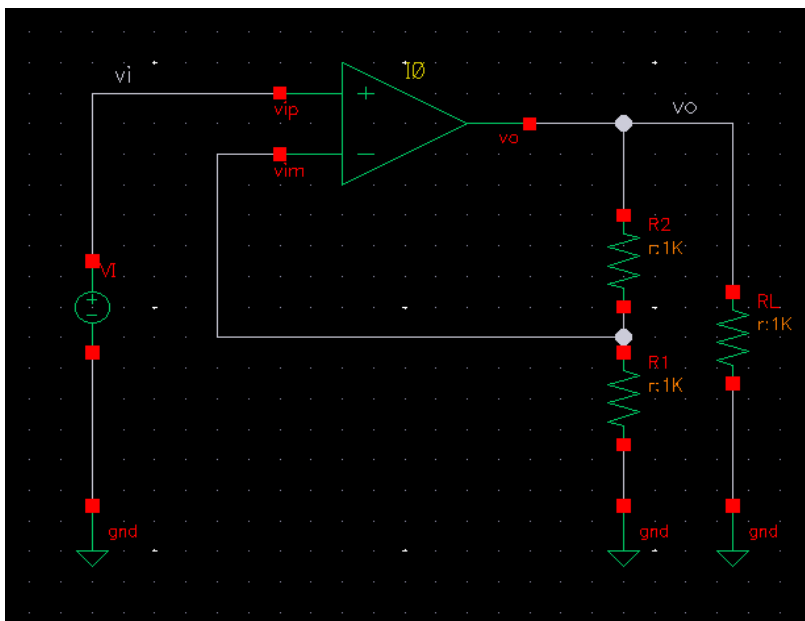
This document is a work in progress and not all sections are complete.

1

Non-inverting amplifier

Open *noninverting_amplifier_1 - config* in the getting_started category (sub-category of examples) of the *gnt* library. Make sure that the view to use of cell *opamp_voltage_1* is 'schematic'. Now click 'open' to open the circuit. This is the well-known opamp in non-inverting configuration. Instance I0 is the operational amplifier. R_L , R_1 and R_2 are the load and feedback resistances, respectively. All resistances are 1 k Ω . The circuit is fed by a symmetrical supply of ± 10 V.

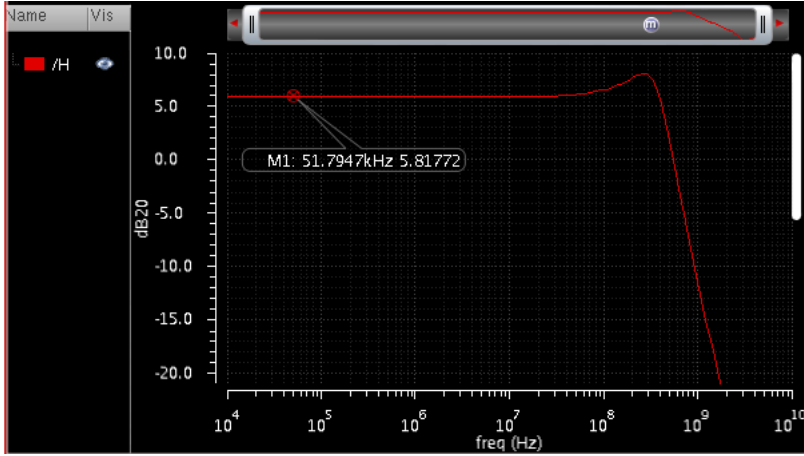
You will learn how to decompose a transfer function of a non-inverting amplifier, with an operational amplifier in non-inverting configuration, using the general feedback theorem (GFT).



Classic theory tells us that the input-output transfer function is approximately given by:

$$H = v_o/v_i \approx 1 + R_1/R_2 = 2 \quad (1.1)$$

The GFT decomposition of H will confirm this. For reference, the simulated H is depicted below. Note that the dc gain is a tad lower than 6 dB, that the system bandwidth is limited and that peaking occurs at higher frequencies. The main goal in this example is learning how to apply the gnt analysis in Virtuoso, along the way learning why the response differs from the ideal.



1.1 Choice of gnt analysis type

The first step is choosing which type of analysis you want to perform. The choices are NEET, 2GFT and 2CT for respectively applying the extra element theorem, general feedback theorem and chain theorem¹. We want to decompose the input-output transfer function H with the classic result Eq. (1.1) as the reference. Indeed, this expresses the ideal, desired response that we, as designers, aim for. The 2GFT decomposes H as:

$$H = H_{\infty} \frac{T}{1+T} + \frac{H_0}{1+T} \quad (1.2)$$

H_{∞} is the response that results when the *loop gain* T becomes infinite ($T \rightarrow \infty$) and is called the *ideal transfer function*. We want it to be exactly $1 + R_2/R_1$. H_0 is the response when the loop gain vanishes ($T \rightarrow 0$) and is called *direct forward transmission*^{2,3}.

Another way of writing Eq. (1.2) is:

$$H = H_{\infty} \frac{1 + \frac{1}{T_n}}{1 + \frac{1}{T}} \quad (1.3)$$

$$= H_{\infty} D D_n \quad (1.4)$$

D and D_n are called the *discrepancy factor* and *null discrepancy factor*, respectively⁴. This is exactly what they are: they express the difference between H and H_{∞} . As H_{∞} is regarded as the ideal, ultimate target, D expresses the changes to H associated with transmission around the feedback loop, in both the forward *and reverse* direction: limited dc gain, limited bandwidth... Likewise, D_n expresses the changes to H associated with direct forward transmission.

1.2 Identification of probe location

A *gnt* analysis decomposes a transfer function using the null injection technique: small-signal test signals (voltage and/or currents) are injected in the circuit, the location of which determines the interpretation of the decomposition result. Hence, the following step is

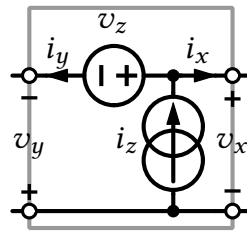
¹ And also NGNT, which returns the raw decomposition, without any meaning attached to the results

² Remark that the GFT emphasizes *loop gain*, not a dependent generator or 'open loop' gain of an amplifier or device. Indeed, the complete loop gain T is what ultimately determines the properties of a loop. GFT starts from the ideal, desired H_{∞} . This contrasts with the similarly looking asymptotic gain model (AGM), which singles out a single dependent generator, but fails to return the desired H_{∞} in the general case. Furthermore, the divide between 'amplifier' and 'feedback path' is artificial or even impossible sometimes. Of course, in idealized circuits, the (2)GFT reduces to the AGM, which is equivalent with the 1GFT.

³ H_0 (or equally T_n or D_n) collects the contributions in the output from all the non-zero signal paths 'in parallel with' H_{∞} , such as forward transmission through the feedback path and common-mode transmission through the forward path.

⁴ $D = T/(1+T)$ and $D_n = 1 + 1/T_n$

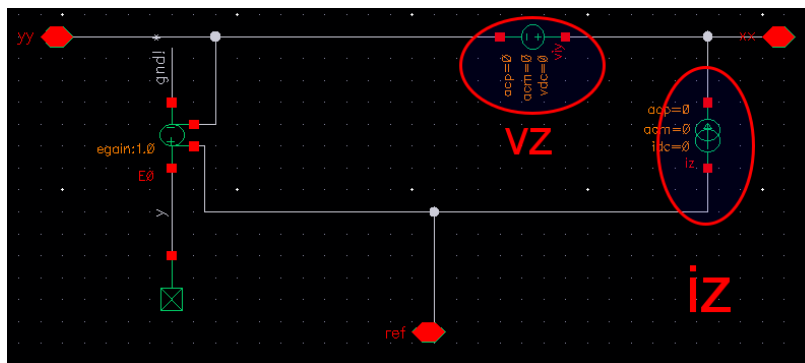
identifying the correct probe location, which defines the test signal injection configuration. A *gntprobe*⁵ contains both a voltage and a current ac source. Their purpose is to inject the test voltage and test current. Two voltages v_x and v_y , as well as two currents i_x and i_y are defined^{6,7}.



⁵ A *gntprobe* performs dual injection at the same location, enabling double-null triple injection calculation. This one is used in a 2GFT or 2CT analysis. A *gntvprobe* or *gntiprobe* perform single voltage or current injection, for null double injection calculations. Those are used in NEET analyses.

⁶ $v_z = v_x + v_y$ and $i_z = i_x + i_y$.

⁷ In the Virtuoso schematic, controlled source E0 does not inject anything, but defines the voltage v_y between terminals *ref* and *yy*. Voltage v_x is the voltage between terminals *xx* and *ref* (note the different directions of v_x and v_y). Current i_y is the current flowing *out* of terminal *yy*, while current i_x is the current flowing *out* of terminal *xx*

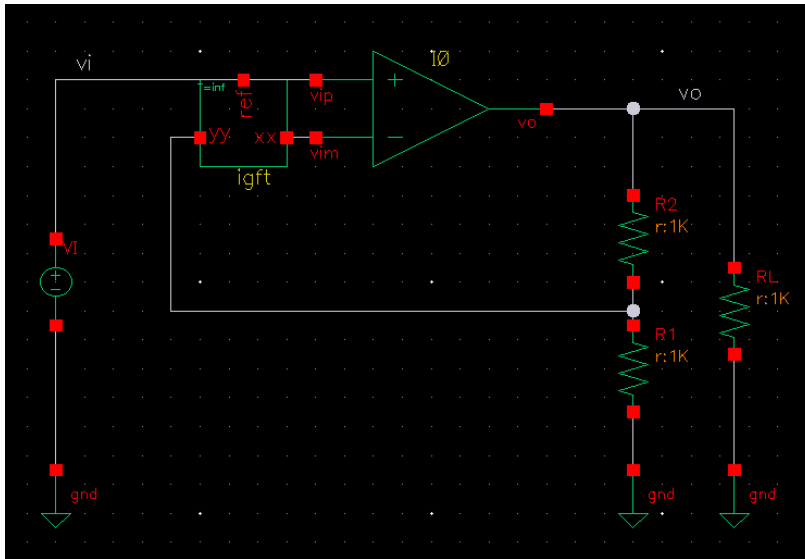


TO CORRECTLY APPLY THE 2GFT, i.e. to obtain the desired H_∞ , *two conditions* must be satisfied: the test signals should be injected 1) in the loop and 2) such that when the total loop error signal (both i_y and v_y) is nulled⁸, H_∞ will assume the desired interpretation. In most circuits, this is the place where you would apply the nullator hypothesis⁹. Note that the location as well as the direction of the probe are important: voltage v_x and i_x should look in the forward direction of the loop, v_y and i_y should look backwards¹⁰. As we want H_∞ to be *exactly* $1 + R_1/R_2$, a *gntprobe* – *symbol* is must be inserted at the following location:

⁸ This is equivalent to simulating infinite loop gain across all frequencies.

⁹ A nullator has zero voltage across *and* zero current through its terminals.

¹⁰ Which direction is ‘forward’ or ‘backwards’ is up to you, the designer. The forward loop is the one that is deliberately shaped and considered by the designer to obtain a given functionality.

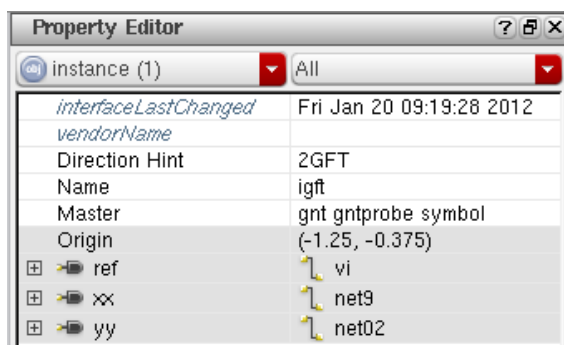


Indeed when both v_y and i_y are nulled¹¹, the voltage between R_1 and R_2 equals the input voltage and H_∞ is the desired expression. The probe is upside-down, to make sure the test voltage is injected in the loop.

¹¹ Nulling a voltage across two nodes is *not* the same as shorting the nodes! Equally, nulling a current in a branch is not the same as open-circuiting the branch.

Direction Hint

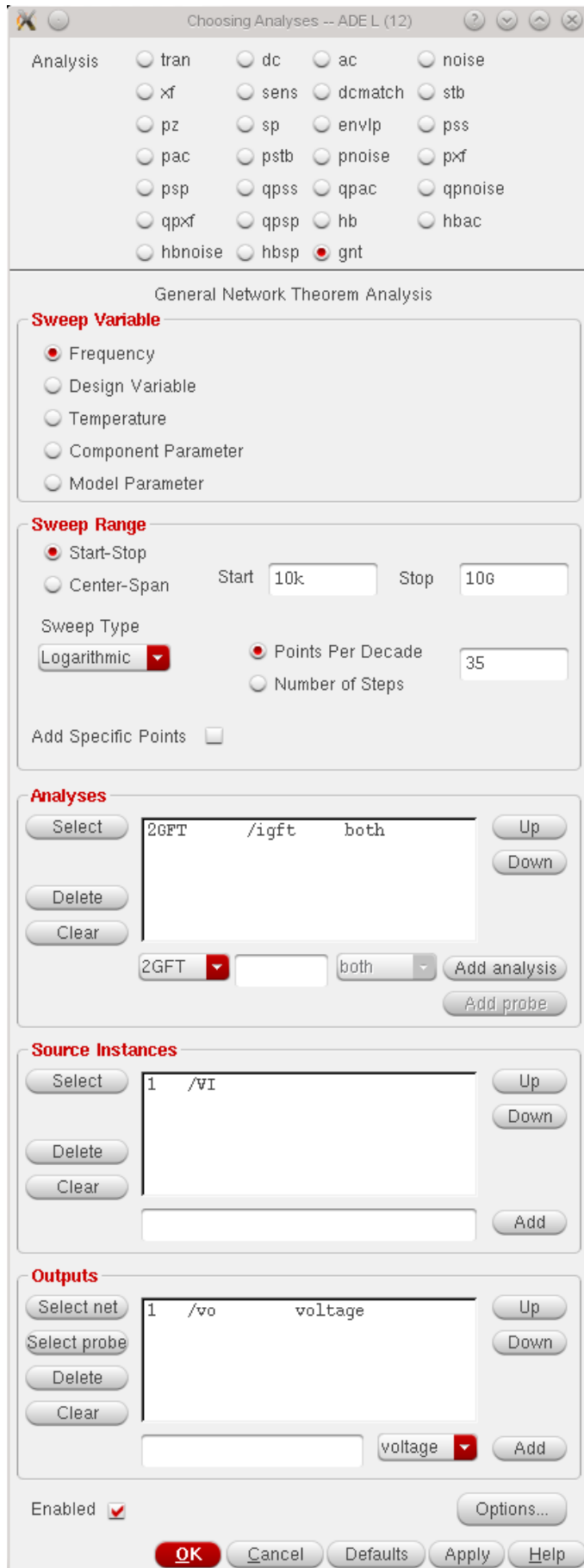
To help you choosing the correct probe location, a hint can be displayed on the probe symbol. Click on the probe, open the Property Editor and select the *Direction Hint* parameter. When '2GFT' is selected, the hint indicates that the location of v_y and i_y must be such that they represent the total loop error signal. In that case, when those signals are nulled, infinite loop gain T is simulated and the desired H_∞ is produced. This is shown above.



1.3 Setting up the GFT analysis

Now that we have identified the probe location, the GFT analysis must be set up. Open the Analog Design Environment (ADE), open the *Choosing Analyses...* form and click *gnt*¹². The following form pops up:

¹² When performing small-signal simulation, it is always good practice to first check that a correct dc operating point is found by running a *dc* analysis (which will be run anyway at the start of any small-signal analysis).



Sweep Variable: defines the parameter you wish to sweep. Usually you want a frequency sweep.

Sweep Range: defines the range across which to sweep the selected Sweep Variable.

Analyses: defines which analyses you want to perform: 2GFT, NEET or 2CT; single or nested. Each analysis is associated to one or more probes you have added to the circuit (*gntprobe*, *gntprobe* or *gntiprobe*).

Source Instances: defines one or more excitation sources—voltage sources or current sources.

Outputs: defines one or more outputs: a voltage is added by selecting a net, a current is selected by adding a probe (*iprobe* or zero voltage source). Together with the Source Instances, this defines all the transfer functions you want to consider (decompose).

Set up the form as shown above:

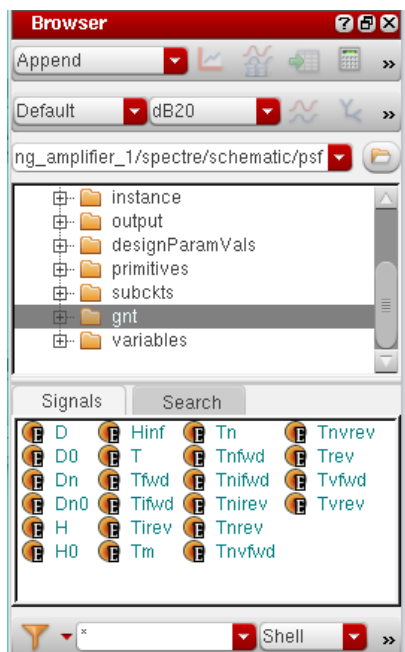
- Enter the Sweep Variable and Sweep Range as in the form.
- Click *Select* in the Analyses frame and select the *gntprobe* in the circuit. Make sure '2GFT' is selected and click *Add analyses*.
- Click *Select* in the Source Instances frame and select voltage source VI. The source is automatically added to the form.
- Click *Select net* in the Outputs frame and select net *vo*. The output net is automatically added to the form.

It is not necessary to define the *AC magnitude* of the sources, as is required for *ac* analyses.

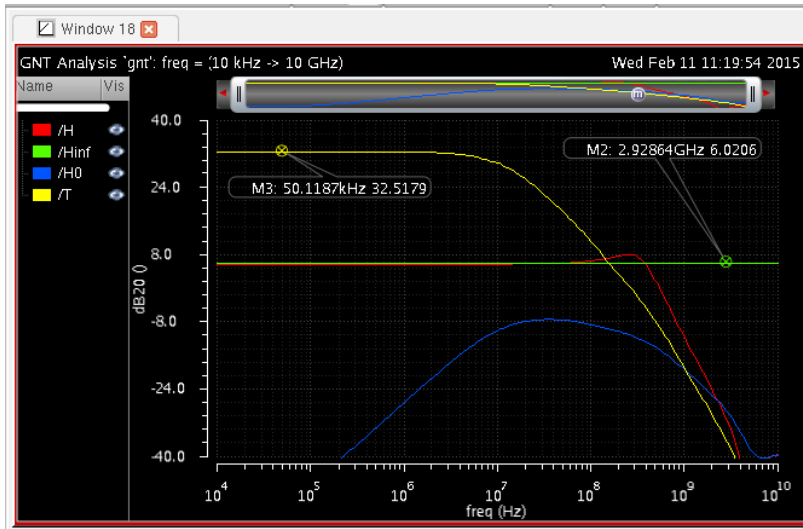
1.4 GFT decomposition of the I/O transfer function H

Run the simulation. You will see some warnings about *alter* statements in the Spectre output log. These can safely be ignored. At the end of the output log some info about the *gnt* simulation is shown. Open the Results Browser to access the simulation results¹³:

¹³ The Direct Plot form is not supported



Select the *gnt* folder to show the 2GFT decomposition result—the lower-level transfer functions. Select the plot modifier *dB20* and double-click 'H', 'Hinf', 'Ho' and 'T'. Adjust the Y-axis limits from -40 dB to 40 dB.

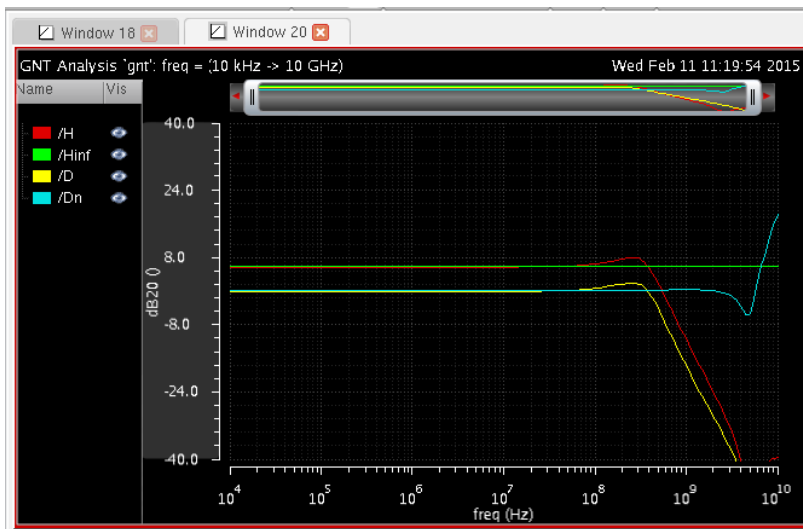


The first thing to do after running a 2GFT simulation is checking whether H_∞ is the desired result, to confirm the probe location. Clearly, this is correct: H_∞ is *exactly* 2, for all frequencies.

Let's investigate the form of the GFT given in Eq. (1.2). We see that H_0 , the direct forward transmission through the feedback path is negligible at all frequencies, also those where the loop gain vanishes. This shows that H_0 does not¹⁴ significantly affect the closed-loop transfer function H . Hence, the frequency domain peaking is not due to direct forward transmission in this case.

¹⁴ This is certainly not always the case.

This is confirmed when we consider the alternative form of the GFT in Eqs. (1.3) and (1.4). Plot 'H', 'Hinf', 'D' and 'Dn' in a new window.

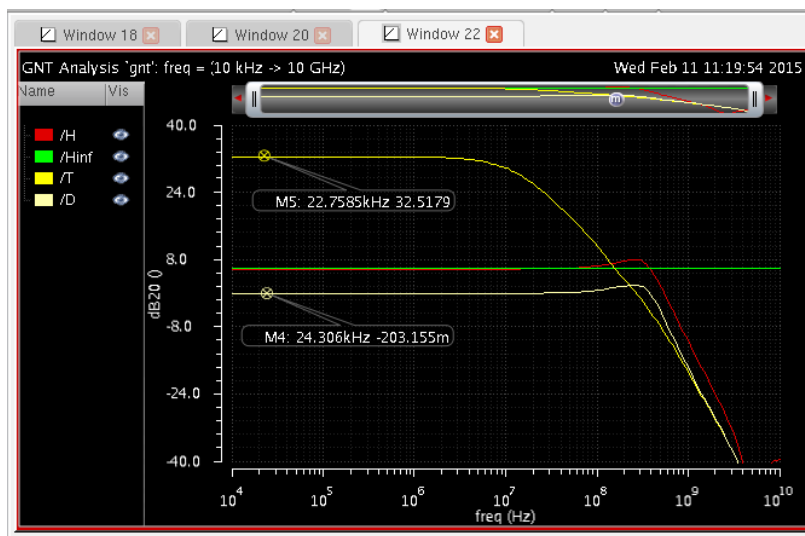


Clearly, the shape of H follows closely that of D . The discrepancy factor D is the ratio of T and $1 + T$, with T the loop gain. This ratio is easily determined using the graphical technique *algebra on the graph*¹⁵. The null-discrepancy factor D_n is 0 dB everywhere, except at very high frequency where it increases. However, this increase has no significant effect as it happens where D is already low.

¹⁵ R.D. Middlebrook. Structured Analog Design, 2004. URL http://rdmiddlebrook.com/D_0A_Rules&Tools/index.asp

LOOP GAIN IS IMPORTANT AS

- it changes¹⁶ the closed-loop transfer function from the ideal H_∞ through D ¹⁷;
- it can be used to assess stability of the closed-loop system^{18,19,20}.



Plot the magnitude of 'H', 'Hinf', 'T' and 'D'. The loop gain T starts out flat with dc gain 32.5 dB. A first pole occurs around 10 MHz. Then the loop gain magnitude assumes a slope of approximately -20 dB per decade. T crosses unity-gain at $f_{\text{odB}} = 270$ MHz²¹. Its slope has already increased further as a second pole occurs around 310 MHz²². The other parasitic poles and zeros of T are at frequencies sufficiently above f_{odB} .

Remark that it is difficult to determine the pole or zero frequencies as the following plot of the slope per decade of T shows. However, it should be clear that the slope starts out at 0 dB per decade, before assuming a slope of around 20 dB per decade, then 40 dB per decade and so on. Sometimes, a plot of the phase enables better determination of the pole and zero locations.

¹⁶ It creates a discrepancy (together with D_n) between what is ideally desired (H_∞) and what eventually results (H).

¹⁷ Loop gain also controls input and output impedances and the sensitivity of H to some circuit elements.

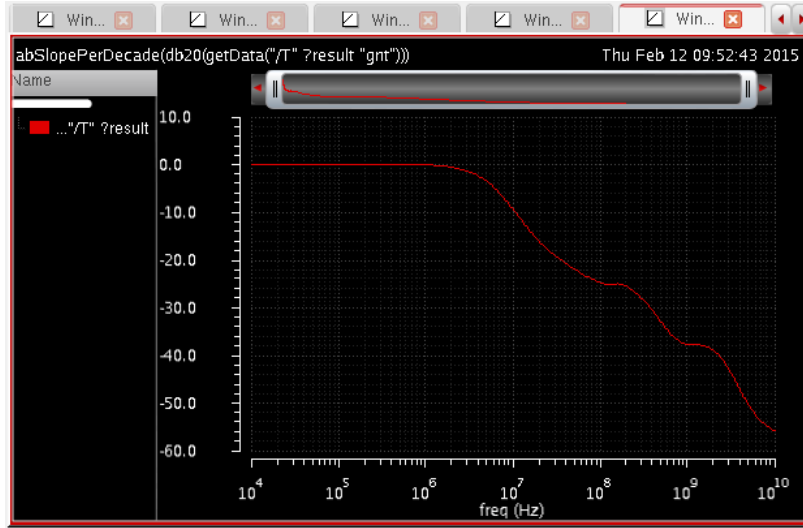
¹⁸ More specifically: of the loop under consideration, assuming that all other loops are stable when this loop vanishes. If desired, multiple-loop feedback theory can be applied using nested 2GFT analyses.

¹⁹ The Nyquist criterion is the ultimate test for stability and can be performed numerically. Other tests for stability, such as the Routh-Hurwitz criterion or Liénard-Chipart criterion are of limited practical value without analytical expressions of the network determinant. In practice, the Bode criterion also gives the correct answer, but only for minimum-phase transfer functions. A transfer function is minimum-phase if its poles and zeros lie within the left-half s -plane.

²⁰ Wai-Kai Chen. *Active network analysis*. World Scientific Publishing Co., Inc., January 1991. ISBN 9971-50-912-1

²¹ Also called the gain-bandwidth product (GBW), which is only formally correct when all parasitic poles and zeros occur at much higher frequency than f_{odB} .

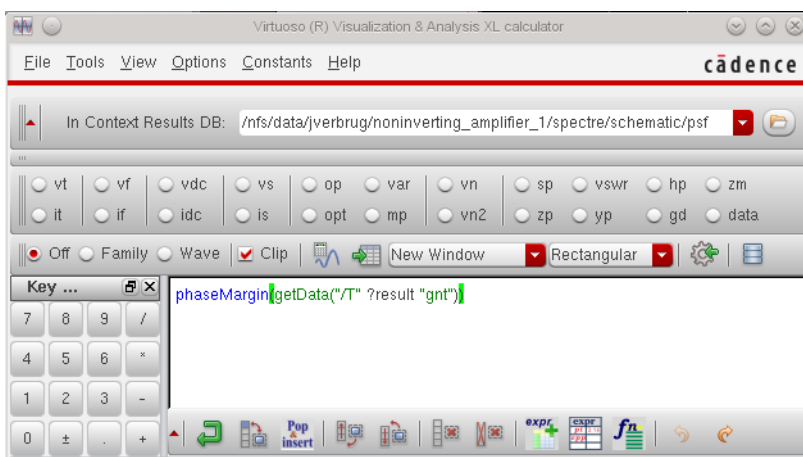
²² This is more easily seen on the phase plot of T



To obtain D , we need to invert the expression $1 + 1/T$. First T is inverted. Then 1 is added. Approximately, this results in a second order system with associated natural frequency (equal to the geometric mean of GBW²³ and the second pole frequency) and quality factor. Finally invert this second order system. This approximative graphical technique is explained in ²⁴ and is highly useful in analytical calculations.

For low frequencies, $D = 1/(1 + 1/T) = 1/(1 + 1/42.27) \approx 0.977 \approx -0.2$ dB. This explains why H is lower than H_∞ ²⁵.

At higher frequencies, the peaking in D is caused by the location of the second pole of T , which is too close to the GBW. Alternatively, one can say that the phase margin of T is too low. Right-click on 'T' in the Results Browser and select Calculator. Replace 'dB20' by 'phaseMargin'²⁶:



The resulting phase margin is 47° , which is quite low and in line with the expectations, because of the peaking in D ²⁷. The system is stable. This can be confirmed on the Nyquist plot of T . In the Results Browser, select the 'RealsvImag' plot type and plot 'T'²⁸. Zoom in on $(-1, 0)$.

²³ The 'true' GBW, even if it does not coincide with f_{odB}

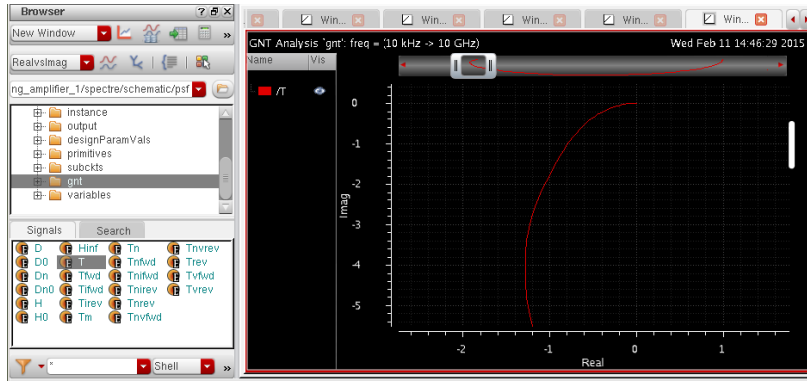
²⁴ R.D. Middlebrook. Structured Analog Design, 2004. URL http://rdmiddlebrook.com/D_0A_Rules&Tools/index.asp

²⁵ This is the static error of the closed-loop transfer function.

²⁶ The sign of the GFT T is the inverse of the stb loopGain. Hence, the phase of GFT T typically starts at 0° , while that of stb loopGain starts at 180° . There are other differences between stb loopGain and GFT T , which are beyond the scope of this document.

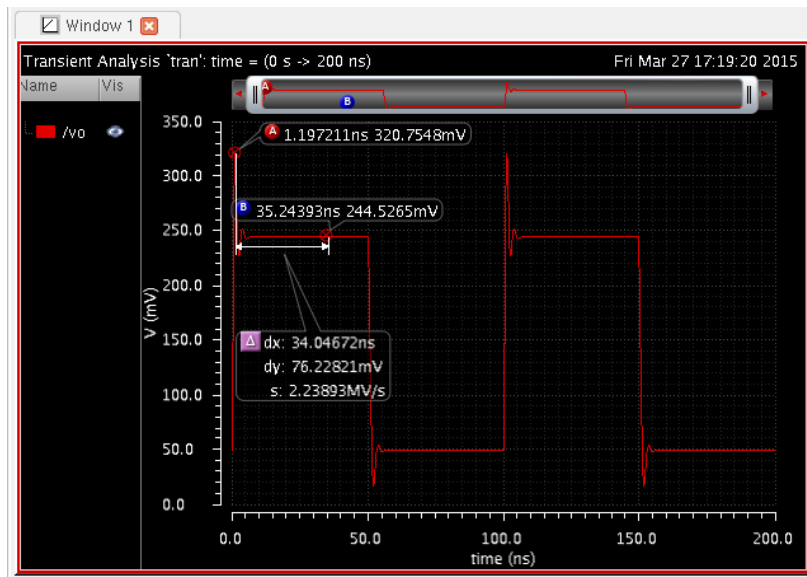
²⁷ Peaking in H can be caused by peaking in D or through non-negligible H_0 . The first case indicates low phase margin of T , the latter *doesn't*.

²⁸ Note that this graph shows the magnitude and phase of T for *positive* frequencies, whereas a real Nyquist plot sweeps over positive and negative frequencies. The curve for negative frequencies is the same as for positive frequencies, but mirrored along the real axis.



The plot does not encircle the critical point, hence the system is stable^{29,30}

In the time-domain, peaking in D will lead to overshoot and ringing³¹. How to setup a transient simulation with step input is left as an exercise to the reader. The following plot shows the response to a 100 mV step input³².



Furthermore, investigation of the loop gain shows that the limited bandwidth of H is caused by the limited bandwidth of D , and hence ultimately by T .

IN SUMMARY, the closed-loop gain of this circuit departs from the ideal in three ways: lower than expected dc gain, peaking and bandwidth limitation. We have shown that, in this case, all these effects are caused by the loop gain. However, not all these effects can be avoided as the loop gain must be shaped—or compensated³³—to make the system stable. Direct forward transmission was found to be negligible.

²⁹ HW Bode. *Network analysis and feedback amplifier design*. D. Van Nostrand Co., Inc., 1945. ISBN 0882752421

³⁰ We assume tacitly that the system that remains when T vanishes is stable. This very important point should not be neglected. Probably the best explanation of the Nyquist criterion is given in Bode's book.

³¹ In contrast, frequency domain peaking due to direct forward transmission typically causes *preshoot* in the time domain.

³² In principle, the input signal should be infinitely small to retain valid small-signal approximation. In practice, a small input that keeps the voltages and currents at all nodes and branches small is sufficient. In the plot, the asymmetric overshoot for rising edges compared to falling edges indicates significant operating point shift and the input signal amplitude should probably be chosen smaller.

³³ Indeed, one compensates a *loop*, not an amplifier!

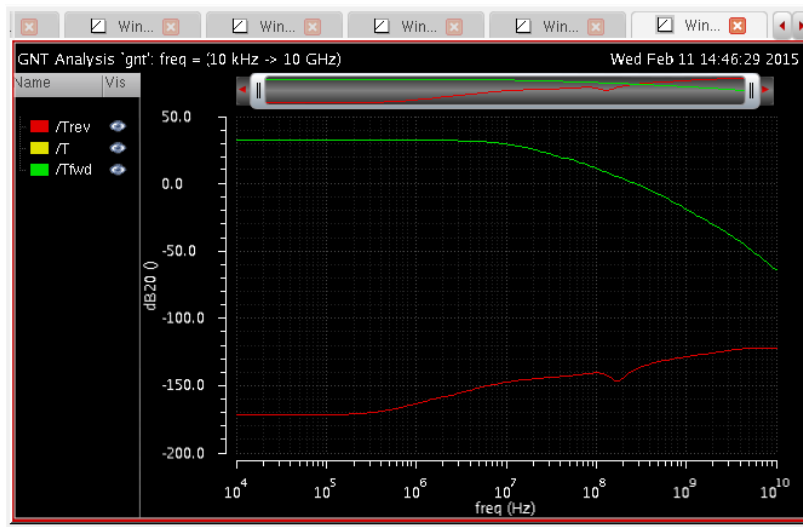
1.5 Digging deeper in the loop gain—improving the circuit

We can now further investigate how the loop gain can be improved to make the closed-loop system better, i.e. that is approximate the ideal transfer function to a better degree³⁴. Let's say we want the low-frequency gain to be closer to 2 and we want to avoid the time-domain overshoot. The previous section already hints at the solution: we want higher dc loop gain and we need to avoid peaking in D . This will come at a cost³⁵: reduced bandwidth. Let's quantify these specification as follows: a static error³⁶ of less than 1% and no peaking.

The GFT loop gain T is given by³⁷:

$$T = \frac{T_{\text{fwd}}}{1 + T_{\text{rev}}} \quad (1.5)$$

T_{fwd} and T_{rev} are the forward and reverse loop gain, respectively. A small T_{rev} indicates a highly unilateral loop, which makes matters considerably easier. Plot the magnitude of 'T', 'Tfwd' and 'Trev':



Clearly, T_{rev} is very small everywhere and $T \approx T_{\text{fwd}}$. Now, T_{fwd} is given by the parallel of the forward open-circuit voltage loop gain and forward short-circuit current loop gain³⁸:

$$T_{\text{fwd}} = T_{\text{v,fwd}} \parallel T_{\text{i,fwd}} \quad (1.6)$$

Plotting the magnitude of 'Tvfwd' and 'Tifwd' reveals:

³⁴ Acceptable tolerances are normally imposed by specifications.

³⁵ Nothing is free in engineering

³⁶ This is a small-signal specification and has *in se* nothing to do with large-signal dc offset.

³⁷ This formula very nicely exposes that we are actually dealing with a multiple-loop feedback system. The reverse loop is just another loop in the system which is considered by the *gntprobe*. Compare with Eq. (1.1).

³⁸ The parallel operator '||' is easily evaluated on the Bode diagram: the smallest one wins.



More dc loop gain

Clearly, at dc, $T_{v,fwd}$ determines T , while at higher frequencies $T_{i,fwd}$ takes over. In fact, the second pole frequency of T is approximately given by the intersection of both curves³⁹.

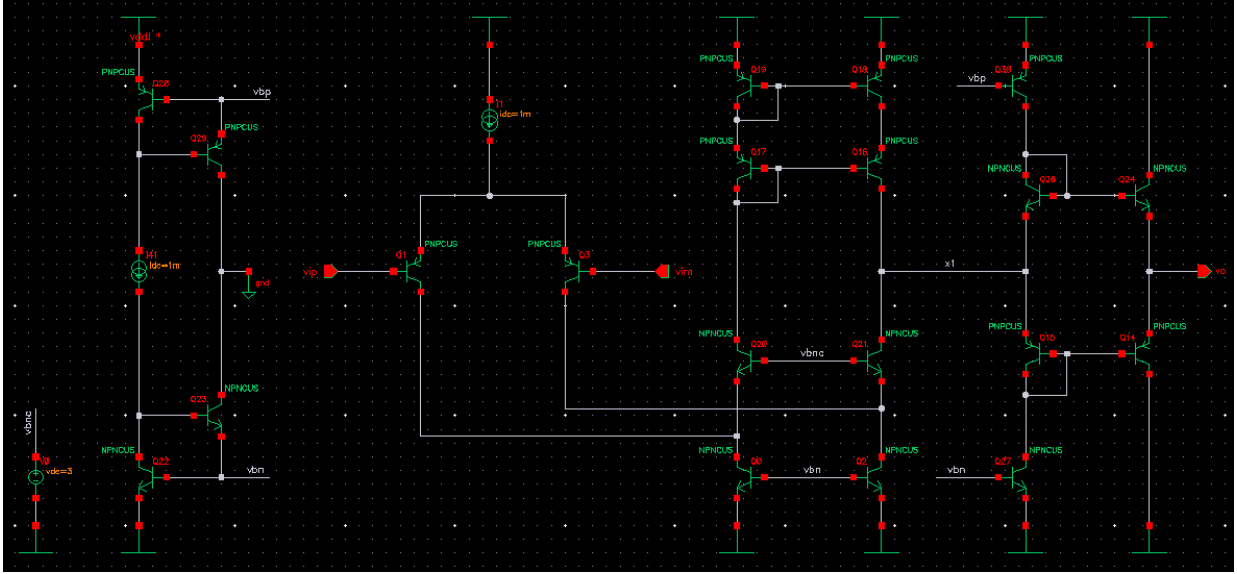
It is instructive to set up a (hopefully) simple analytical expression for the dc loop gain. This serves three purposes: it explains the simulation result in terms of circuit elements, might provide a way to improve it and confirms whether your understanding of the circuit is correct.

Remark that the opamp does not function in isolation. Its behavior is always influenced by the external circuit, although buffers might mitigate these effects. As such, asking for the gain of the opamp has no meaning or stipulating its phase margin makes no sense; in feedback systems *the loop gain* is an important quantity, which is exactly what the GFT emphasizes.

Descend into the opamp. This is a bipolar single-stage⁴⁰ amplifier with class AB output buffer. The input stage is differential pair with folded cascode load. As mentioned above, the purpose of the buffer is to decouple the high impedance inner node x_1 from the low resistive external load (and also enables high output currents).

³⁹ In fact, if we would put the probe at an ideal injection point in the circuit, only $T_{v,fwd}$ or $T_{i,fwd}$ would remain, yet still the same T would appear. Indeed, as long as the probe is in the (same) loop in the correct direction, the same T results. Unfortunately, there is no such point that is accessible as is often the case in real circuits. Furthermore, if T_{rev} is not negligible, H_∞ would *not* be the same.

⁴⁰ Here, 'stages' of an amplifier are voltage amplification stages and create high-impedant internal nodes.



We already know from simulation that only the forward voltage loop gain $T_{v, \text{fwd}}$ is important at dc. A simple model of the dc loop gain is therefore given by⁴¹:

$$T_0 \approx g_{m1} (R_{o1} \parallel R_{i2}) \frac{R_1}{R_1 + R_2} \quad (1.7)$$

in which g_{m1} , R_{o1} and R_{i2} is the average transconductance of the input bipolar transistors, output resistance of the cascode and input resistance of class AB buffer, respectively. R_{o1} is the parallel of the output resistances (βr_o) of the upper and lower cascode branches and is very high compared to R_{i2} . R_{i2} is the parallel input resistances of the upper and lower branches of the bias circuit of the output transistors and is approximately equal to the output resistance of rail devices. Hence $R_{o1} \parallel R_{i2} \approx R_{i2} \approx r_{o30} \parallel r_{o27}$. Looking at the dc operating point of the devices gives us the numerical values:

$$T_0 \approx 17 \text{ mS} (9.4 \text{ k}\Omega \parallel 25 \text{ k}\Omega) \frac{1 \text{ k}\Omega}{1 \text{ k}\Omega + 1 \text{ k}\Omega} \quad (1.8)$$

$$= 58 \approx 35.3 \text{ dB} \quad (1.9)$$

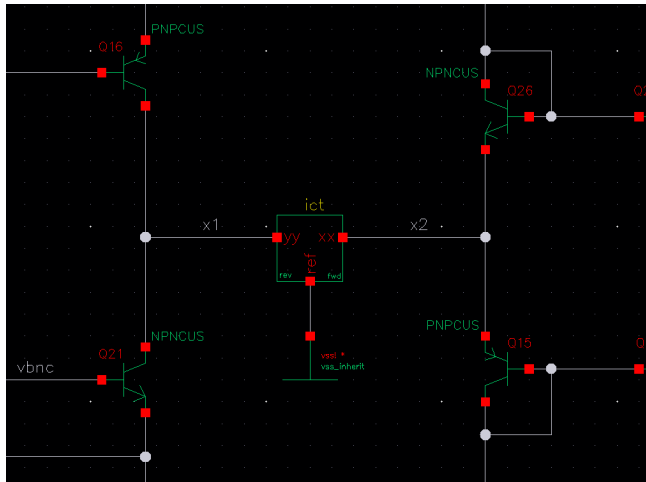
This agrees reasonably well with the simulated value of 32.5 dB.

A static error⁴² of less than 1% calls for a dc loop gain of at least 40 dB. The analytical expression and the reasoning behind provides clues for improving T_0 . We could increase the bias current of the input pair to increase g_{m1} , decrease the loading of the buffer on the cascode or change the external resistances. Clearly, the ratio of the external resistances must remain the same to retain the same H_∞ . Increasing g_{m1} is possible, but more can be gained by reducing the load of the buffer on the cascodes.

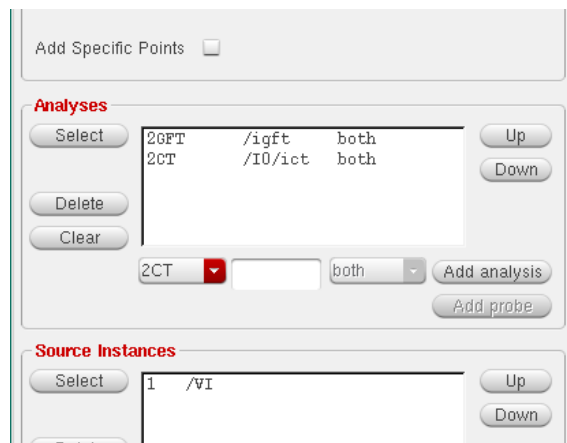
In order to simulate the loading effect and the possible improvement to the loop gain, we will apply the chain theorem (CT) to the loop gain by nesting a 2CT analysis in the 2GFT analysis. Descend into the opamp and insert *gntprobe - schematic* between the cascode stage and the output buffer. To help identifying to correct probe direction, select '2CT' for the *Direction Hint* parameter of the probe:

⁴¹ This is *really* simplified: a.o. , the following is ignored: effective g_{m1} by degeneration, input loss of bipolar devices (base resistance, input resistance), low beta, voltage gain of buffer stage. However, it will be seen to be already sufficiently accurate.

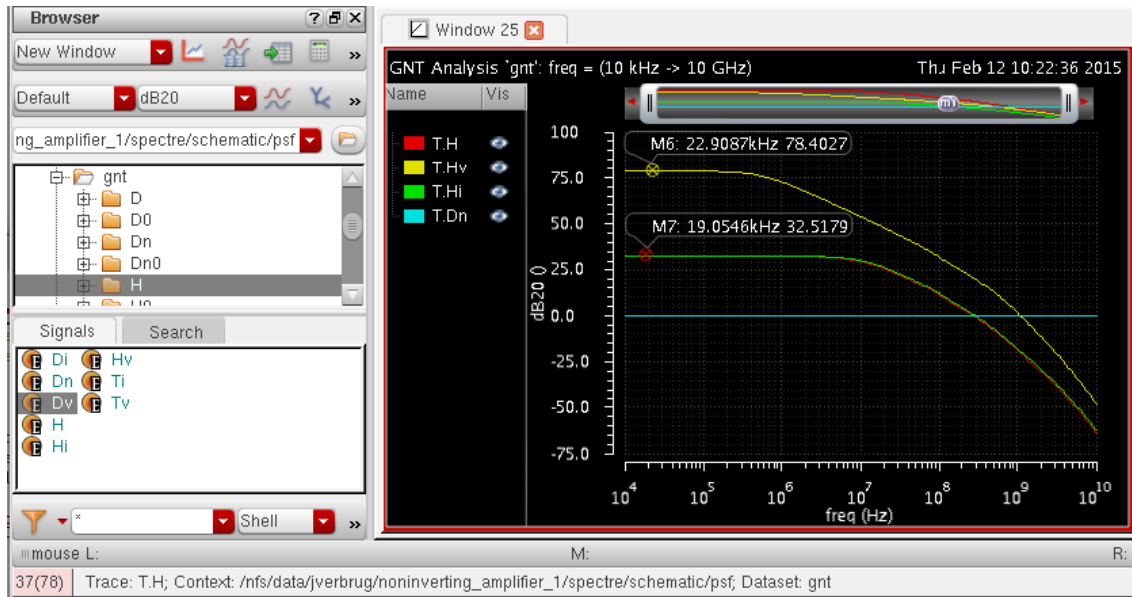
⁴² This error is referred to H_∞ and assumes H_0 zero. Again, this is a small-signal specification, assumes the circuit remains linear, and has generally nothing to do with dc offset.



Terminal *yy* must look in reverse direction, while terminal *xx* must look in the forward direction. Open the *gnt* analysis form, click the *Select* button in the *Analyses* frame and select the probe in the schematic. Make sure '2CT' is selected for the analysis type and click *Add analysis*. The form should look like this:



After running the simulation, open the *gnt* folder in the Results Browser. Note that each result of the 2GFT has been subjected to the 2CT and that the nested results appear as subfolders:

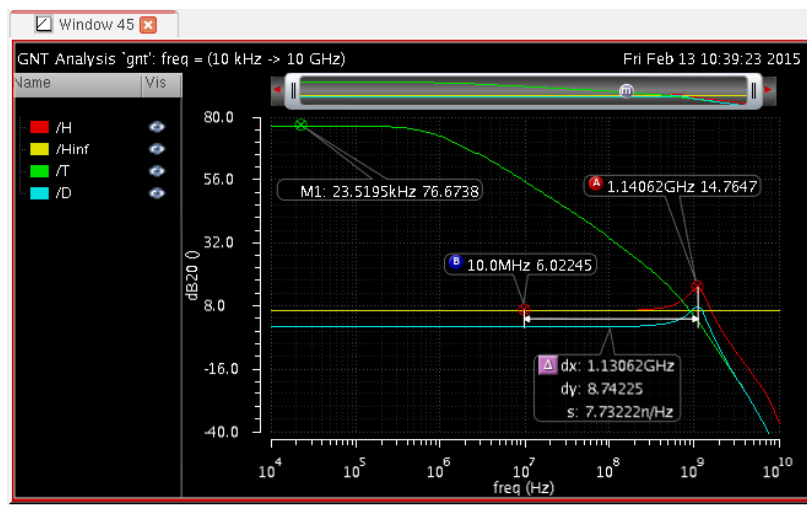


Confirm that $T.D_n$ is close to 0 dB everywhere⁴⁶. The plot of $T.H_i$ shows that we gain nothing by including a current buffer. Indeed, the first stage with its cascode load already works quite well as current source. On the other hand, the $T.H_v$ plot shows that we could potentially gain up to 45 dB dc loop gain if a voltage buffer would be inserted at the probe location. This will be, however, at the expense of stability, which will be solved later.

⁴⁶ This is not true for e.g. H ; indeed the 2CT probe does not break all parallel paths from input to output.

There are various ways to implement this voltage buffering. Here, for the sake of simplicity, we will use an ideal one (a vcvs). In the hierarchy editor, change the view to use for *opamp_voltage_1* to *schematic_voltage_buffer* and save the configuration. Re-descend into the opamp and confirm that a voltage-controlled voltage source is inserted.

Now open ADE and set up a GFT analysis, with the same parameters as in the previous section. Plot 'H', 'Hinf', 'T', 'D':

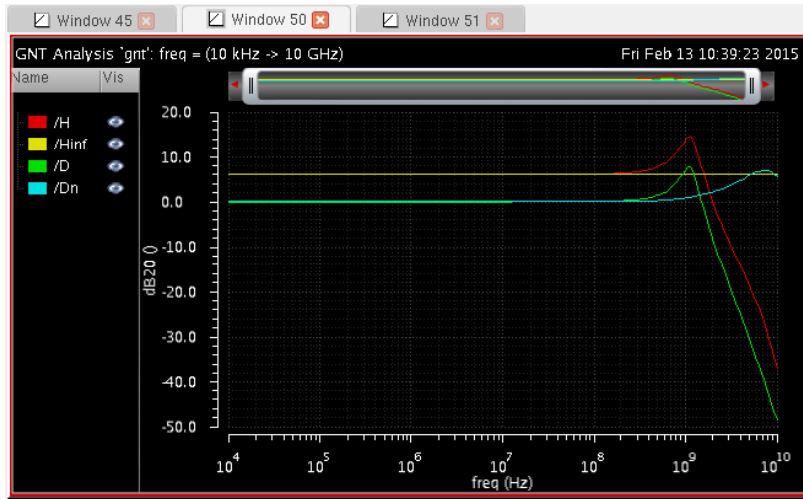


The dc loop gain has improved to 76.7 dB⁴⁷. H_∞ is still exactly

⁴⁷ A bit lower than the predicted 2CT value of 78.4 dB as the inserted voltage buffer changes the dc operating point a bit.

2. However, D and hence H exhibit largely increased peaking, up to 8.7 dB. This indicates a potential stability problem, which will be investigated shortly.

First, let's check whether the direct forward transmission has changed. In a new window, plot ' H ', ' H_{inf} ', ' D ' and ' D_n ':

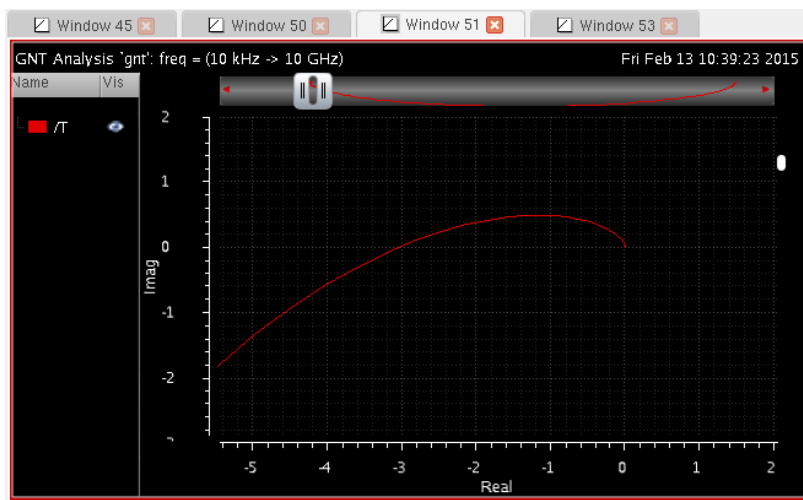


D_n is still close to 0 dB nearly everywhere, and hence has not significant impact on H .

Less peaking

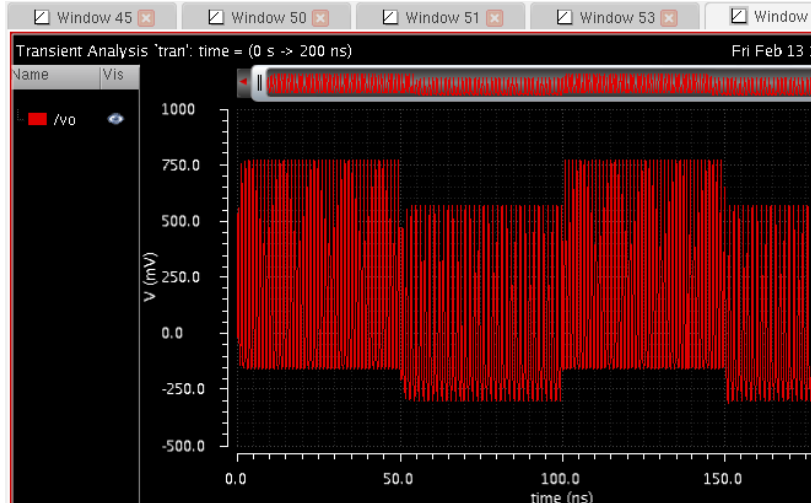
In the previous section we have improved the dc loop gain 100-ish times. However, this has increased peaking in D and H . In this section we will tackle this new issue.

The phase margin of T yields -28° (using the Calculator). This indicates an unstable system. Make the Nyquist plot of T :



The curve encircles the critical point $(-1, 0)$ (clockwise) which confirms that the circuit is unstable. This is also confirmed by a transient simulation. The system oscillates⁴⁸:

⁴⁸ The oscillation is superposed on the step input.



Re-investigation of the magnitude plot of T shows that its second pole appears before the GBW⁴⁹, such that D is approximately a second order system with high Q-factor. As T and hence D contain more than 2 poles, the circuit oscillates⁵⁰. One solution is to bring the GBW down by lowering the first pole frequency. Inspection of the opamp reveals that no explicit compensation capacitance is included. In the following section, the effect of parallel frequency compensation will be investigated using the extra element theorem (EET) analysis.

The practical value of the EET⁵¹ is that it tells you what the effect of an element (or combination of elements) is on the behavior of your circuit. The EET decomposes a transfer function as⁵²:

$$H = H_{\text{ref}} D D_n \quad (1.12)$$

$$= H_{\text{ref}} \frac{1 + \frac{1}{T_n}}{1 + \frac{1}{T}} \quad (1.13)$$

in which H_{ref} is the transfer function when a designated extra element (typically an immittance) assumes either a zero or infinite value (open or short in case of immittances), the latter which is called the reference value. The reference value is determined by the type and direction of a single test signal (voltage or current). However, in most cases you do not need to think about the test signal injection, as the symbol hint on the probe will tell how to place the probe.

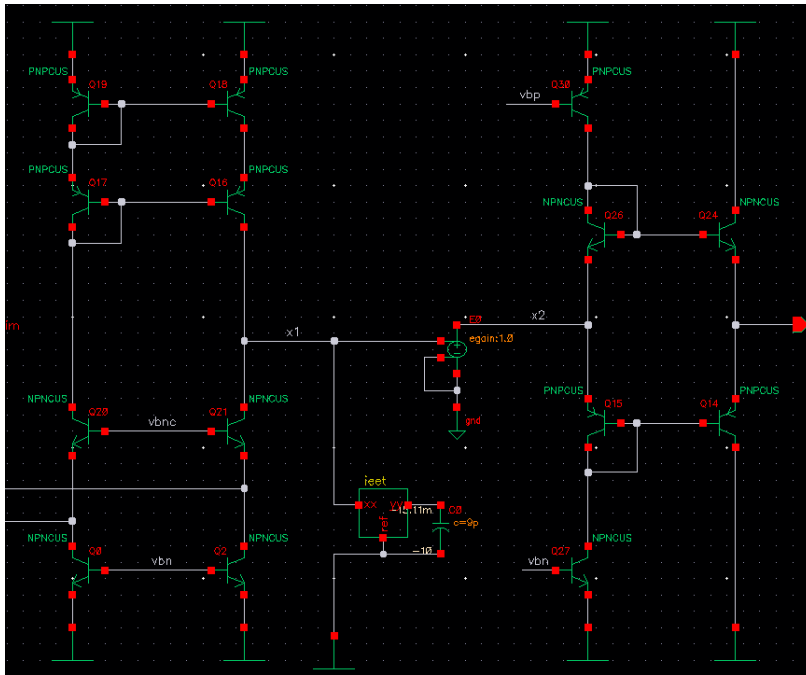
In the hierarchy editor, change the view to use for `opamp_voltage_1` to `schematic_voltage_buffer_comp` and save the configuration:

⁴⁹ Not f_{odB} , but literally the gain \times bandwidth frequency, which is easily identified by extrapolating the part of T which has a 20 dB per decade slope.

⁵⁰ A pure second order system cannot be unstable, although can approach instability infinitely close.

⁵¹ An augmented version for N extra elements (or combinations), NEET gives the effect of multiple elements.

⁵² The EET can also be calculated in terms of (null) driving point immittances.



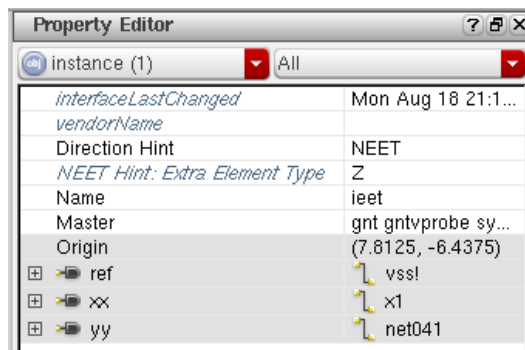
A 9pF compensation capacitance C_L has been added to the high impedant internal output node. Also, an additional probe *gntvprobe*⁵³ has been inserted to analyze the effect of C_L on H and its GFT decomposition using the EET analysis. Assuming the forward voltage loop gain $T_{v, fwd}$ still dominates the loop gain T , a simple expression for the expected GBW can be deduced. If the second pole is high enough, this is approximately equal to f_{odB} :

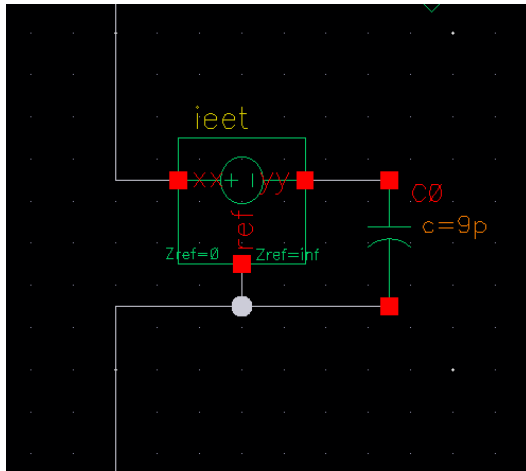
$$f_{odB} \approx \frac{g_{m1}}{2\pi C_L} \frac{R_1}{R_1 + R_2} \tag{1.14}$$

$$\approx 150 \text{ MHz} \tag{1.15}$$

⁵³ For general impedances, the EET can be calculated using either voltage or current injection (*gntvprobe* or *gntiprobe* instances). For dependent sources, only one of both signal types can be used.

The symbol hint on the *gntvprobe* symbol indicates the correct direction of the probe. For the symbol property 'NEET Hint: Extra Element Type', an impedance 'Z' is selected. Then if the capacitance faces the hint 'Zref = inf', the reference transfer functions will be calculated using an infinite reference impedance; as if the capacitance was open-circuited.





Set up a nested NEET analysis in the existing 2GFT analysis. The *gnt* analysis form should look like:

General Network Theorem Analysis

Sweep Variable

- Frequency
- Design Variable
- Temperature
- Component Parameter
- Model Parameter

Sweep Range

- Start-Stop
- Center-Span

Start: 10k Stop: 10G

Sweep Type

- Logarithmic
- Points Per Decade: 85
- Number of Steps:

Add Specific Points

Analyses

Select	2GFT	/igft	both	Up
Delete	NEET	/I0/ieet	voltage	Down

Clear

NEET voltage Add analysis

Add probe

Source Instances

Select	1	/VI	Up
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Delete

Clear

Add

Outputs

Select net	1	/vo	voltage	Up
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Select probe

Delete

Clear

voltage Add

Run the simulation. The nested results, i.e. the EET decomposition of all the results of the GFT, appear as subfolders in the Result

Browser. First we'll check the loop gain. Plot 'T.Href', 'T.H', 'T.D' and 'T.Dn'. Obviously $T.H_{ref}$ is the loop gain that we got in the previous section, when C_L was absent. $T.H$ is loop gain including the extra element. The plots show how the loop gain is affected by the extra capacitance: the first pole frequency has decreased ($T.D$).



Plot T, Tvfwd and Tifwd to confirm the assumption that T is still dominated by $T_{v, fwd}$:

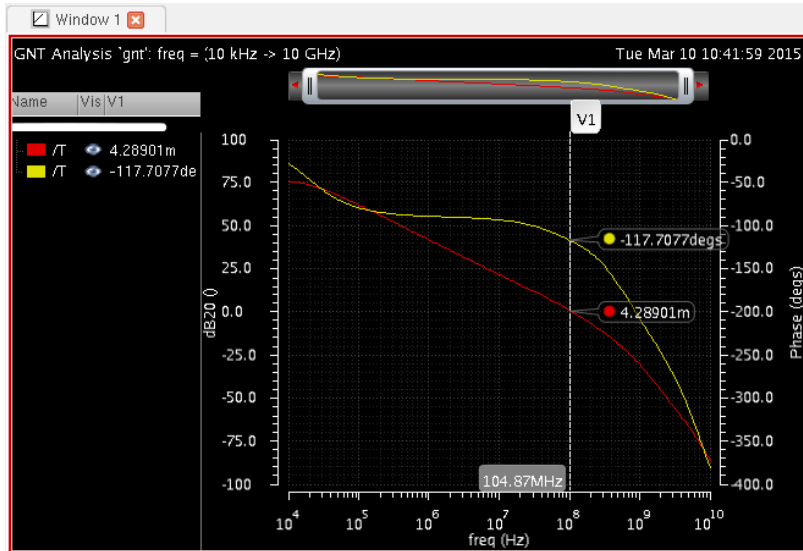


This seems to be the case for low frequencies. After comparing the GBW simulation result with the prediction Eq. (1.15), it is estimated too high. If you, as a designer, decide it is off too much for your purposes, the model must be improved. One enhancement that can

be readily included is the parasitic capacitance at the internal node. This, however, is outside the scope of this exercise.

We can check stability on the Nyquist plot or calculate the phase margin of T using the Bode stability criterion⁵⁴. Let's determine phase margin graphically this time. Plot the magnitude and phase of T :

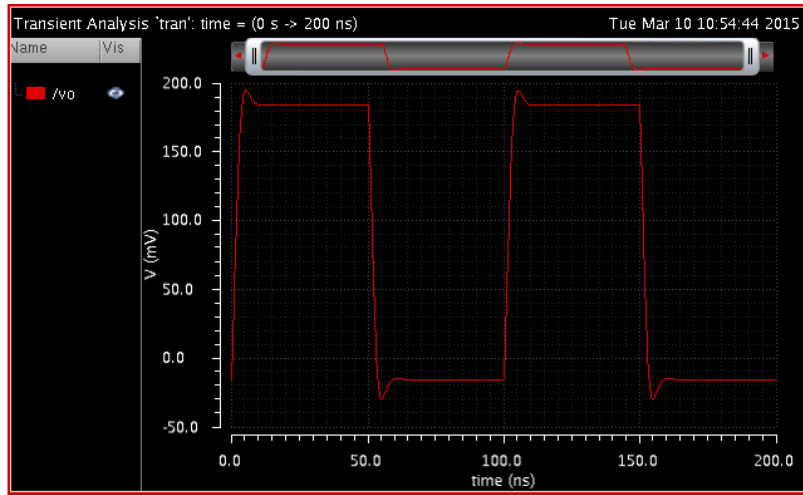
⁵⁴ Recall the limitations of this method.



A vertical marker at the unity-gain frequency returns a phase of -117° . This gives a phase margin of 63° . The circuit is stable. Plot 'H', 'Hinf', 'T' and 'D':



In conclusion: the phase margin gives rise to minor peaking in D and hence in H . The resulting transient overshoot is largely diminished, but at the expense of closed-loop bandwidth. This is confirmed in the transient step response:



2

Multiple-input/multiple-output functionality: analyzing input and output immittance, power supply rejection at once

TBD

3

Multiple-loop feedback circuits

TBD

4

Bibliography

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