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Fundamentals of Logarithmic Amplifiers and Recent Advances

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ANALOG DEVICES

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LOGARITHMIC AMPLIFIERS

ANALOG DEVICES

- Unique Nonlinear Function
- Integrated Multistage Systems
- Calibrated Slope and Intercept
- Provide Complete Solutions
  -- easy to use
- Up to 100 dB Dynamic Range
- Now covering DC - 30 GHz
- Limiter Versions for PSK, FSK
- Low Cost, Small Packages
- Numerous types available

AD606	AD8313
AD607	AD8314
AD608	AD8315
AD640	AD8316
AD641	AD8317
AD8302	AD8318
AD8306	AD8319
AD8307	AD8362
AD8309	AD8363
AD8310	AD8364
AD8311	more…..

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A Personal Goal: Make Log Amps as Cheap, and Easy to Use, as Op Amps

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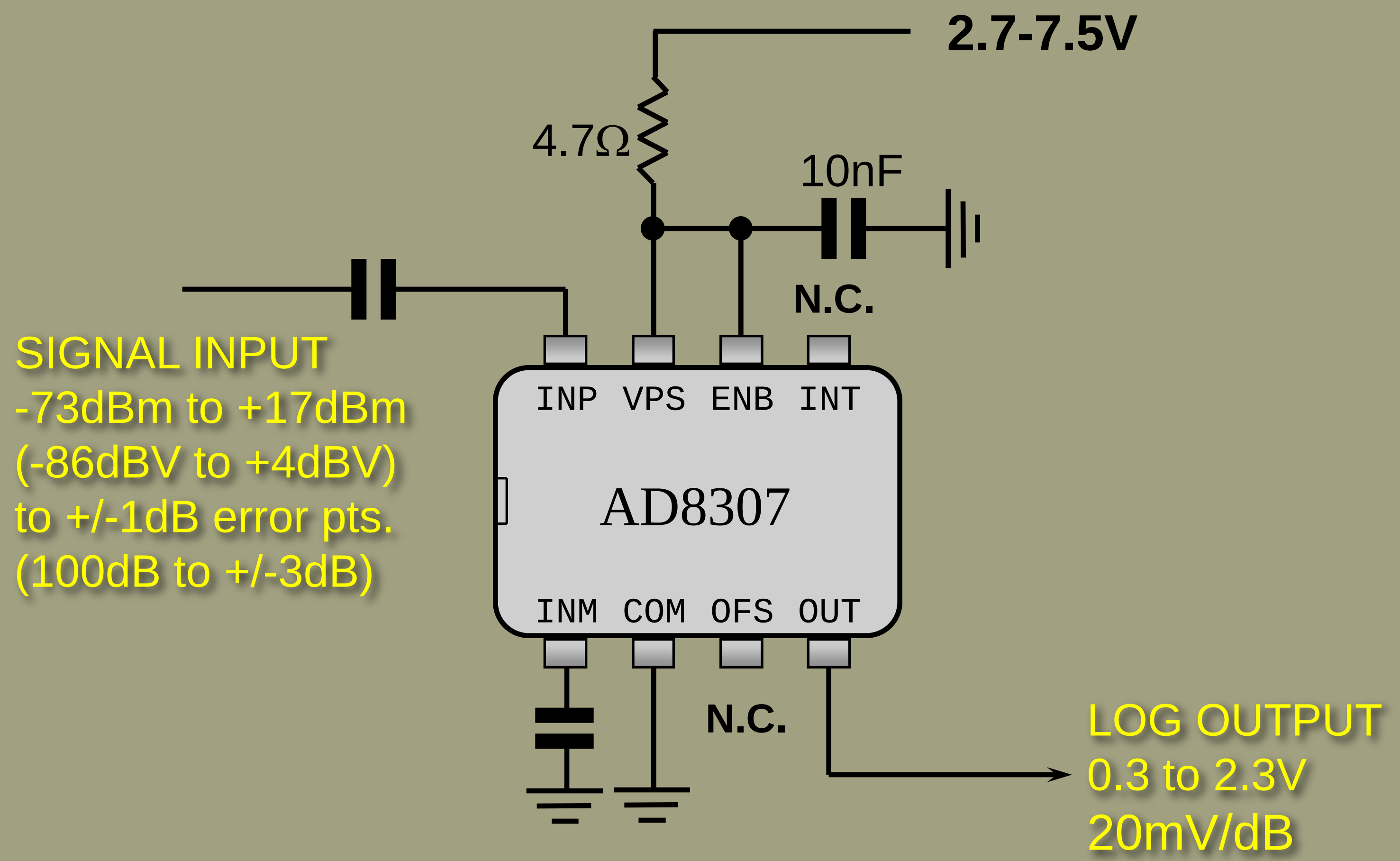

SO HERE'S THE PLAN....

- BEGIN WITH A BRIEF OVERVIEW OF THE VARIOUS TYPES, TO SET THE STAGE
- MOVE ON AS QUICKLY AS POSSIBLE TO PROGRESSIVE COMPRESSING TYPES as likely to be of most value in beam instr.
- DEVELOP THEIR FUNDAMENTAL THEORY starting from the most basic of foundations
- SHOW SOME PRACTICAL EMBODIMENTS

WHAT DO LOG AMPS DO?

- Convert signals of high dynamic range (HDR) to a substantially smaller dynamic range
- The output is readily scaled to represent the *decibel value* of the input, in simple units
- This is a *fundamental nonlinear conversion* of the signal representation - with important consequences
- Some types may be used to simply *compress* a HDR signal, thereby achieving high observational or measurement-sensitivity near a nominal null

FUNDAMENTAL FUNCTION

$$W = Y \log \frac{X}{Z}$$

where

W *is the* Output variable

X . . . . . Input variable

Y . . . . . Slope parameter

Z . . . . . Log Intercept

FUNDAMENTAL FUNCTION

$$V_W = V_Y \log \frac{V_X}{V_Z}$$

where

V_W *is the* Output voltage

V_X . . . . . Input voltage

V_Y . . . . . Slope voltage

V_Z . . . . . Intercept voltage

FUNDAMENTAL FUNCTION

$$I_W = I_Y \log \frac{I_X}{I_Z}$$

where

I_W *is the* Output current

I_X . . . . . Input current

I_Y . . . . . Slope current

I_Z . . . . . Intercept current

FUNDAMENTAL FUNCTION

$$V_W = V_Y \log \frac{I_X}{I_Z}$$

where

V_W *is the* Output voltage

I_X . . . . . Input current

V_Y . . . . . Slope voltage

I_Z . . . . . Intercept current

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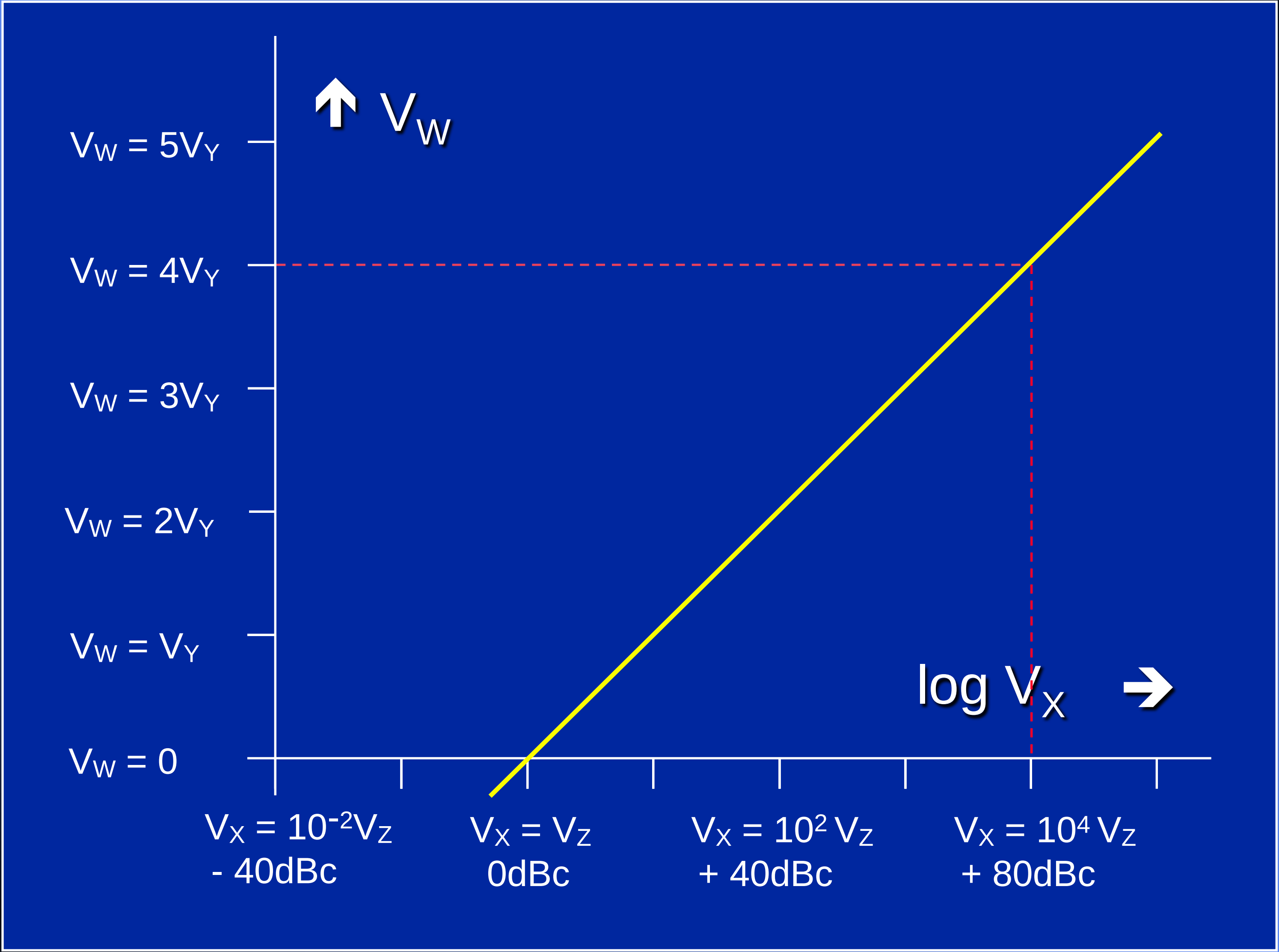
THE BASIC LOGARITHMIC RELATIONSHIP



REGION NEAR ZERO

$$\frac{\partial V_W}{\partial V_X} = \frac{\partial}{\partial V_X} V_Y \left(\log V_X + \log V_Z \right)$$

$$= \frac{V_Y}{V_X}$$

THE INCREMENTAL GAIN OF A LOG-AMP SHOULD APPROACH INFINITY AS $V_X \rightarrow 0$

WHAT HAPPENS WHEN $V_X < 0$?

Formally, log (-X) IS COMPLEX

One might consider using

$$V_W = \mathrm{sgn}(V_X)\, V_Y \log(|V_X| / V_Z)$$

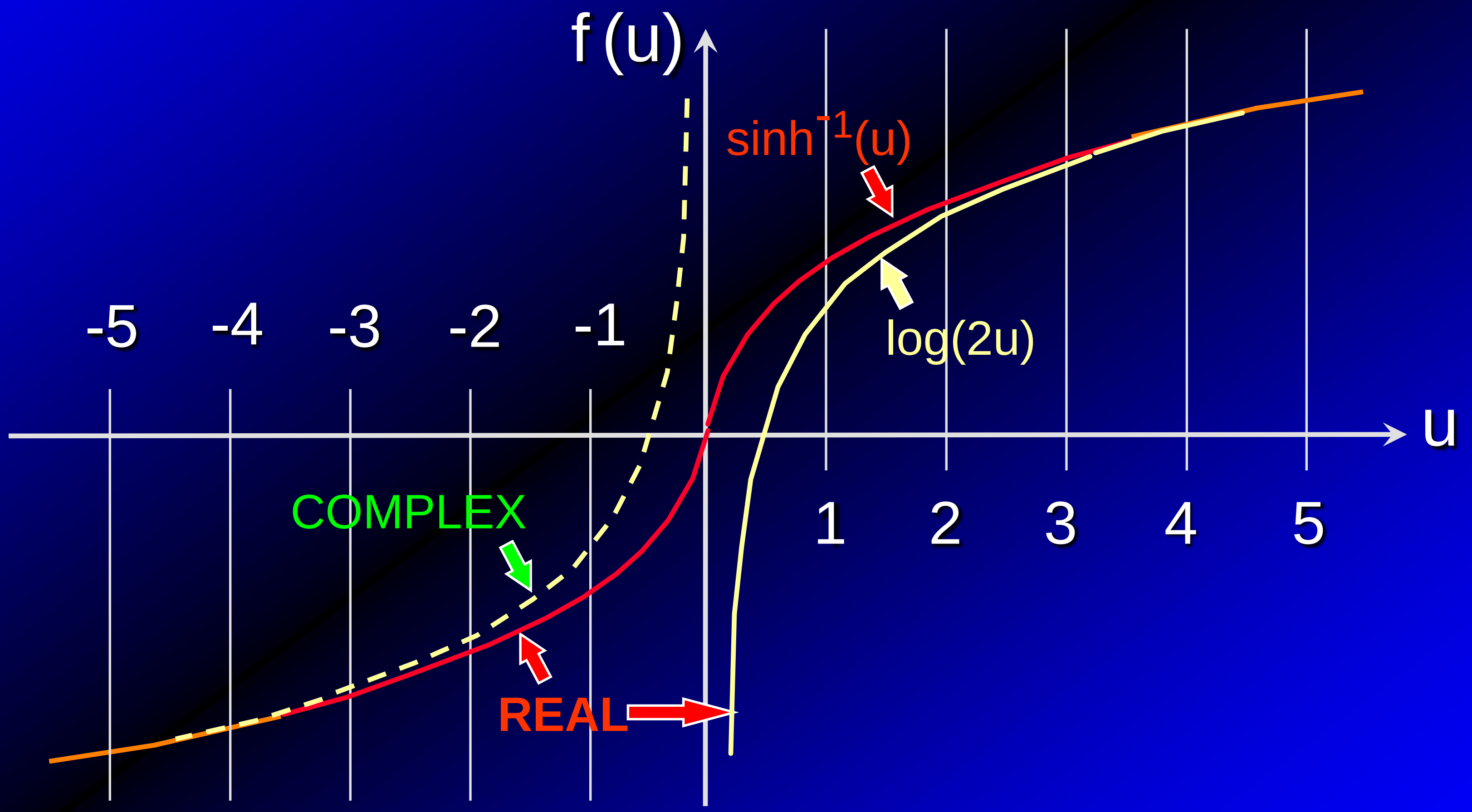
The inverse hyperbolic sine is useful here:

$$\sinh^{-1}(u) = \log\{u + \sqrt{(u^2 + 1)}\}$$

$$\rightarrow \log(2u) \quad \text{for } u > \text{than} \sim 3$$

Of course, $\sinh^{-1}(u) = -\sinh^{-1}(-u)$

$\sinh^{-1}(u)$ & $\log(2u)$



PRACTICAL LOG AMP TYPES

- TRANSLINEAR (inc. DC/LF log-ratio)
- EXPONENTIAL-AGC
- PROGRESSIVE COMPRESSION
  - Baseband
  - Demodulating
  - “True-log” ($sinh^{-1}$)
  - Synchronous
  - Dual (RF log-ratio)

DIRECT TRANSLINEAR

- BASED ON THE LOG-EXPONENTIAL PROPERTIES OF THE BJT
- CURRENT-MODE DYNAMIC RANGE CAN BE VERY HIGH (e.g., 200dB, 1pA to 1mA)
- INVALUABLE IN MANY TYPES OF LF INSTRUMENTATION (DC to about 10MHz)
- THE MOST ADVANCED FORMS ACTUALLY PROVIDE FULL LOG-RATIO OPERATION

THE TOUCHSTONE:

V_{BE}

A widely used formulation is

$$V_{BE}(T, I_C) = \frac{kT}{q} \log \frac{I_C}{I_S(T)}$$

.. which is very nearly exact in most respects, over a temperature range of -250°C to 250°C and an I_C range from (typically) 1pA to 1mA; at high currents the junction resistances and other effects cause V_{BE} to exceed this value. But what is this peculiar quantity $I_S(T)$?

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The Saturation Current $I_S(T)$

If the wondrous V_{BE} can be called the heart of a bipolar transistor, $I_S(T)$ must surely be its soul! Although very tiny, it is also a marvel to behold, and derives from the expression for the intrinsic carrier concentration – the number of free holes and electrons per unit volume generated by the *thermal energy* in an unbiased semiconductor:

$$n_i^2(T) = 32\left(\pi \frac{k}{\hbar} m_e m_h\right)^3 T^3 \exp\frac{\alpha}{k} \exp\frac{-E_{GO}}{kT}$$

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The Saturation Current $I_S(T)$

In practice $I_S(T)$ cannot be accurately calculated -or even measured directly - for use in the basic expression $V_{BE} = (kT/q) \log I_C / I_S(T)$.

Instead, the V_{BE} of a representative transistor is measured at a known temperature and current; then a different formulation for V_{BE} is used for calculating its value at other operating points.

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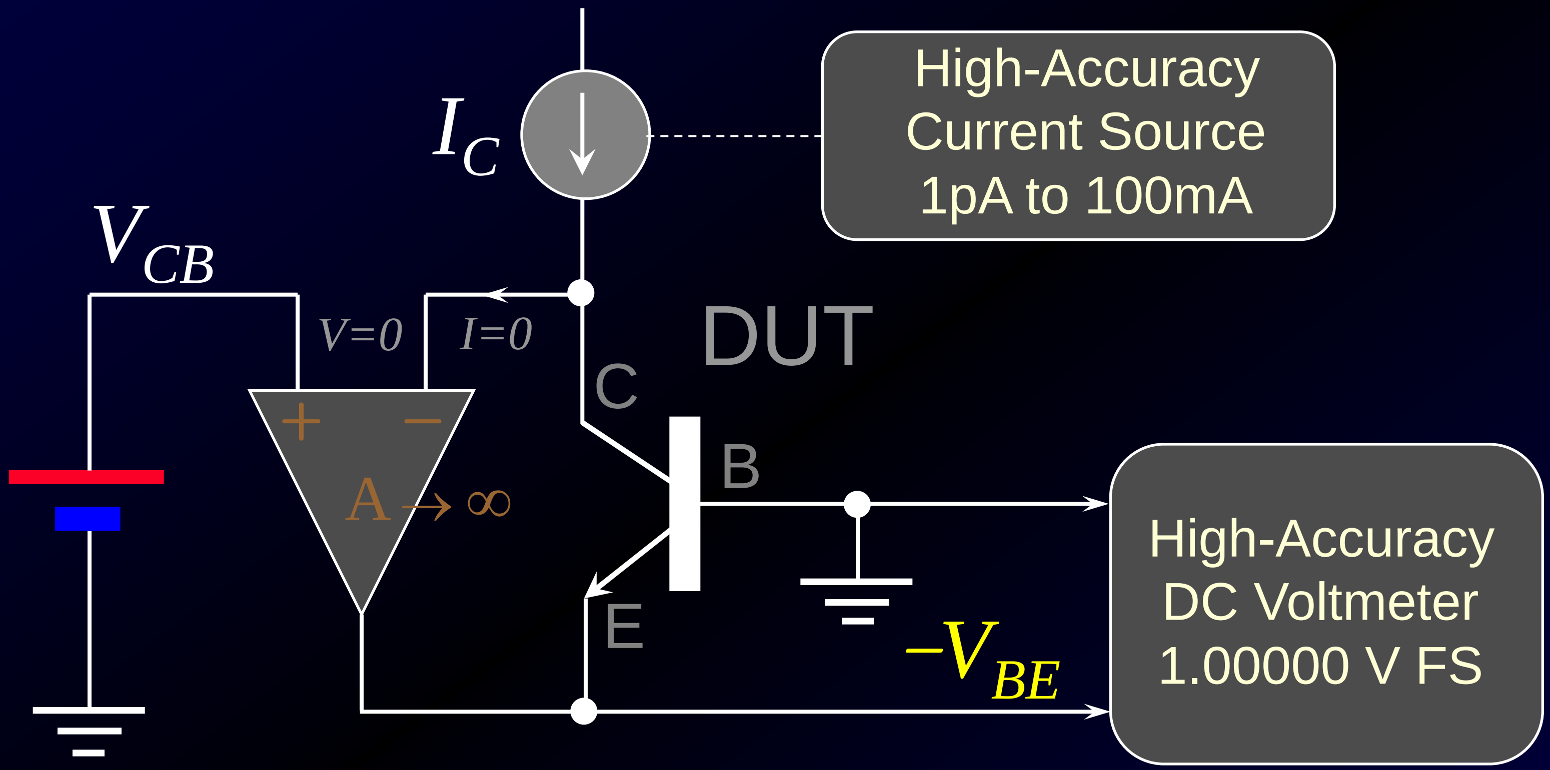
Measurement of $V_{BE}(T, I_C)$

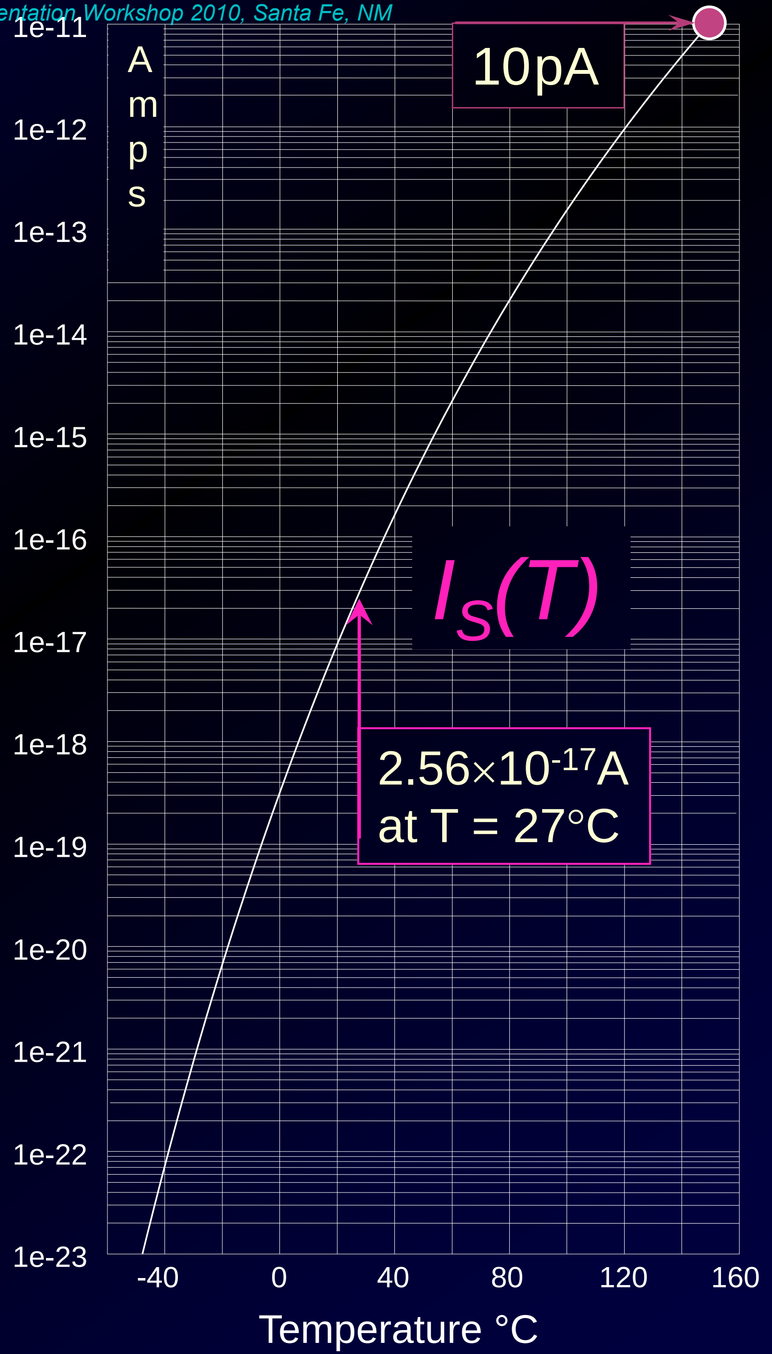


In BJT modeling, the default collector bias is $V_{CB} = 0$, that is, $V_{CE} = V_{BE}$. Collector current I_C is forced by an electrometer-grade op amp

The form

$$V_{BE}(T,I_C) = \frac{kT}{q} \log \frac{I_C}{I_S(T)}$$

might suggest that V_{BE} has a strong positive coefficient of temperature - because of the factor kT/q. However, it is the <u>saturation current</u> $I_S(T)$ which dominates, climbing by a ratio of about a <u>trillion</u> from -50°C to +150°C. Graph is for a V_{BE} of 0.75V at I_C = 100uA & 300K



 Mead @ Lausanne, July 2003

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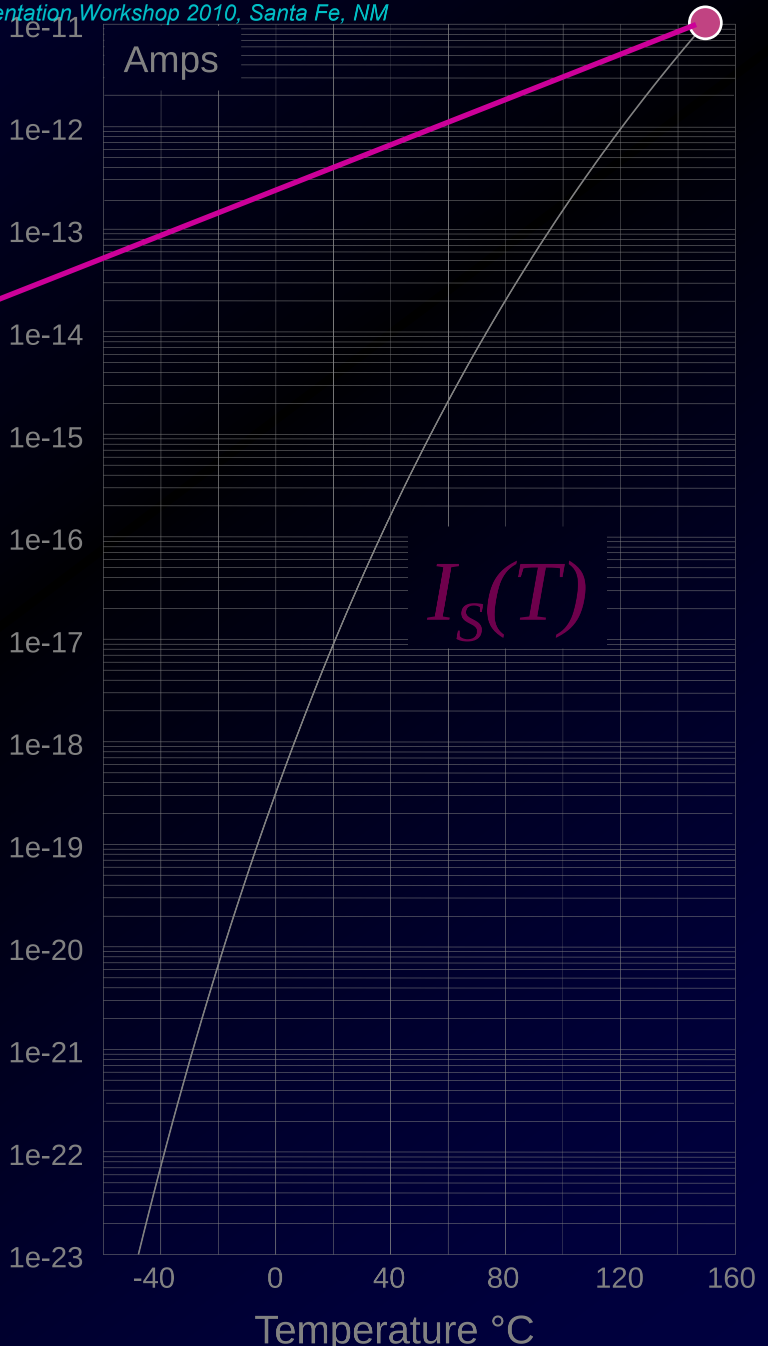
$$V_{BE}(T, I_C) = \frac{kT}{q} \log \frac{10\,\text{pA}}{10\,\text{pA}}$$

also suggests that the V_{BE} of this transistor would become ZERO at 150°C for I_C = 10pA …. which indeed really does happen, because all of I_C is then supplied entirely by I_S! In fact, at high temperatures and picoamp current levels, V_{BE} may become negative !!



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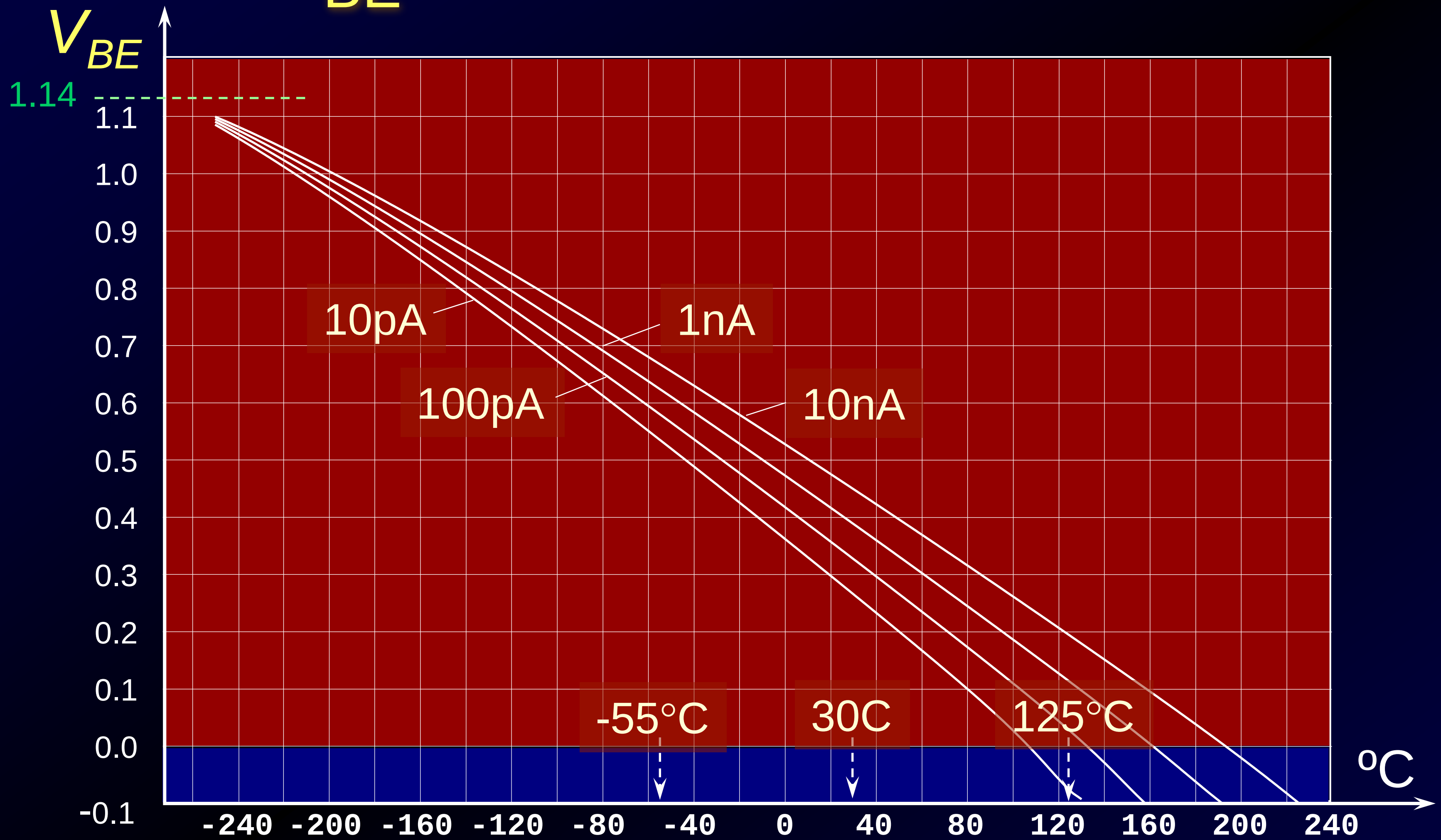
V_{BE} VS. TEMPERATURE



Actual PNP transistor having LE=100um, WE=2um, EG=1.13, XTI=4.03, IS = 2.644e-16

I_C = 10pA, 100pA, 1nA and 10nA. To optimally illustrate the effect, VBC was adjusted to 52mV (10pA), 83mV (100pA), 110mV (1nA) and 150mV (10nA) - all values ZTAT

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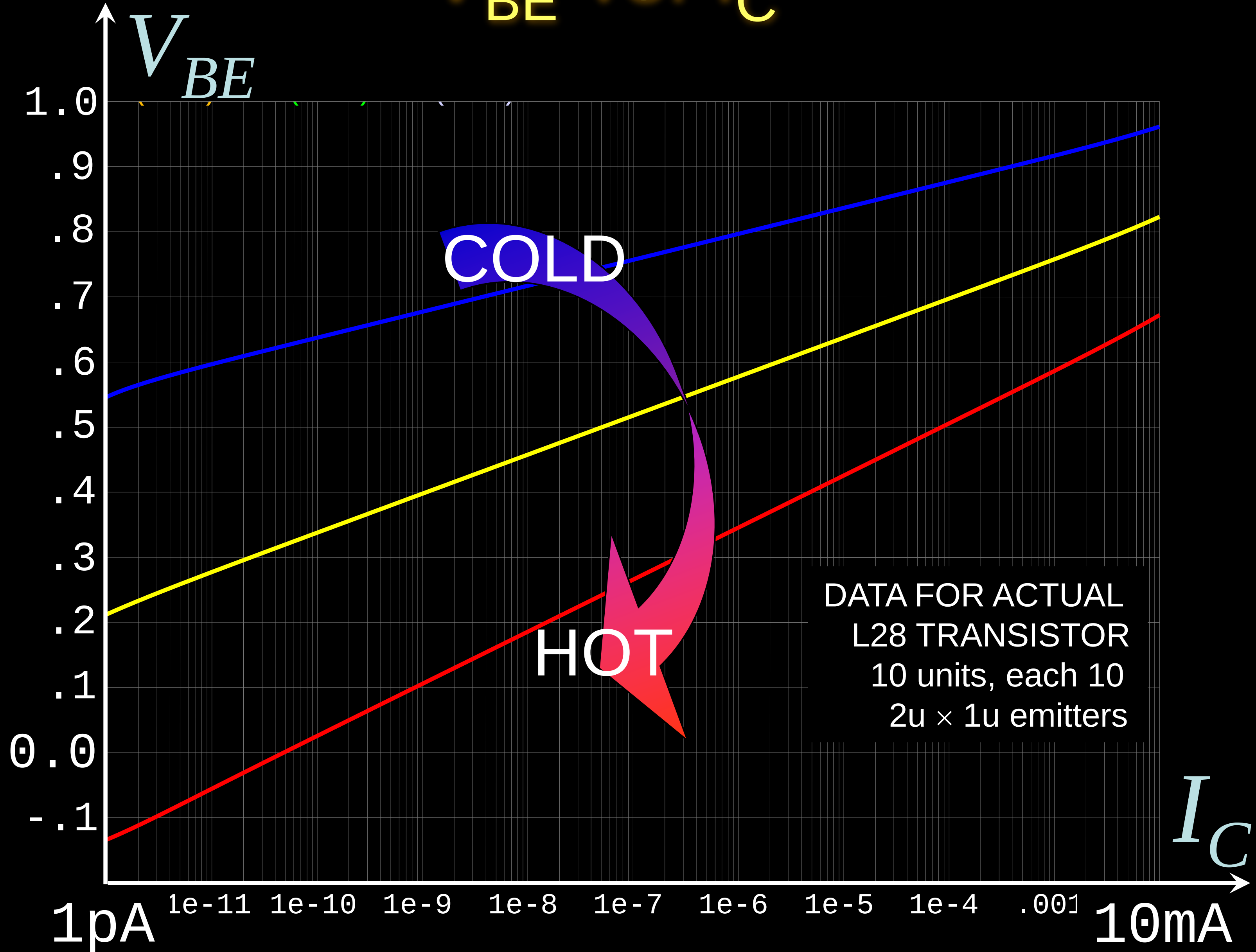
V_{BE} vs. I_C



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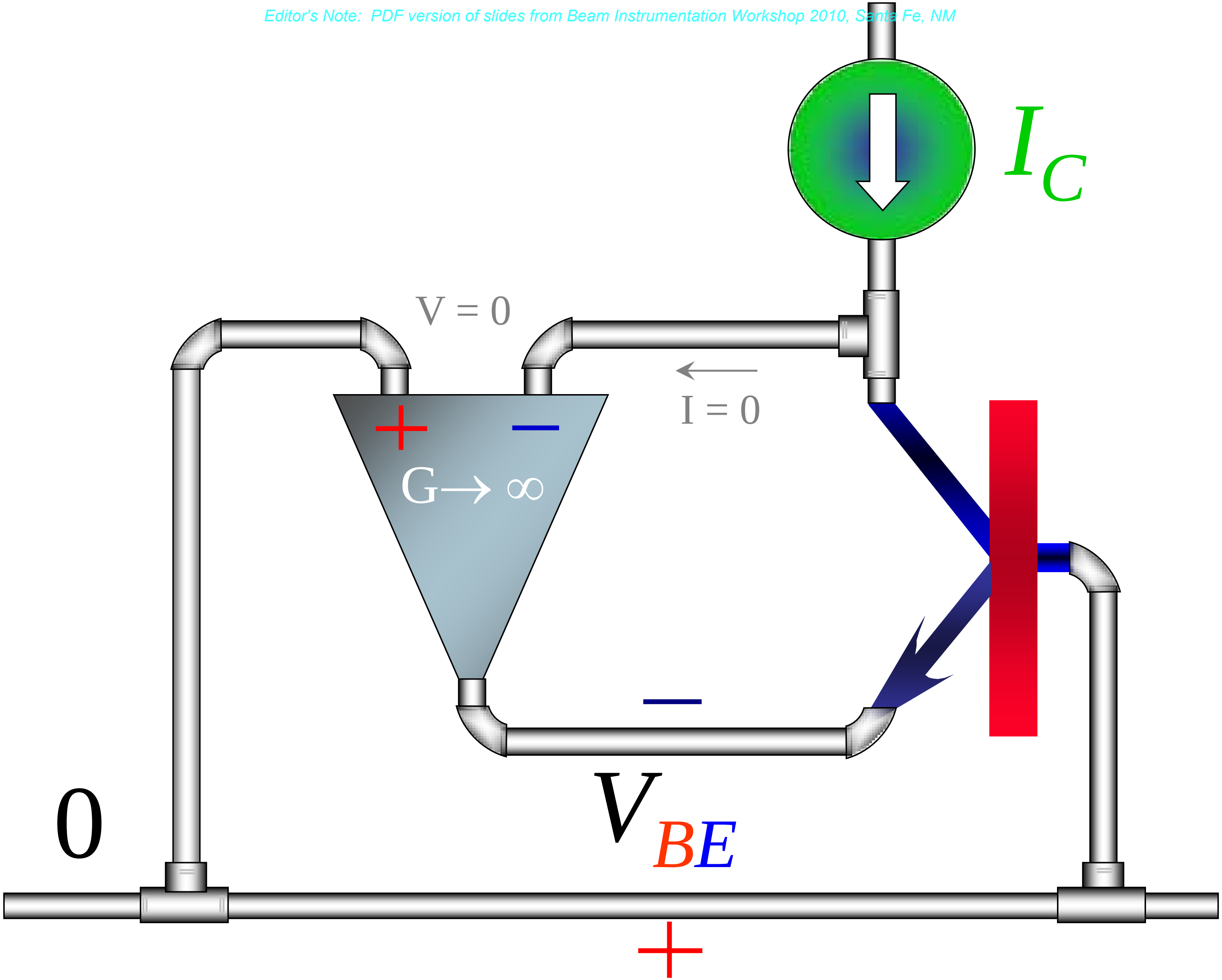
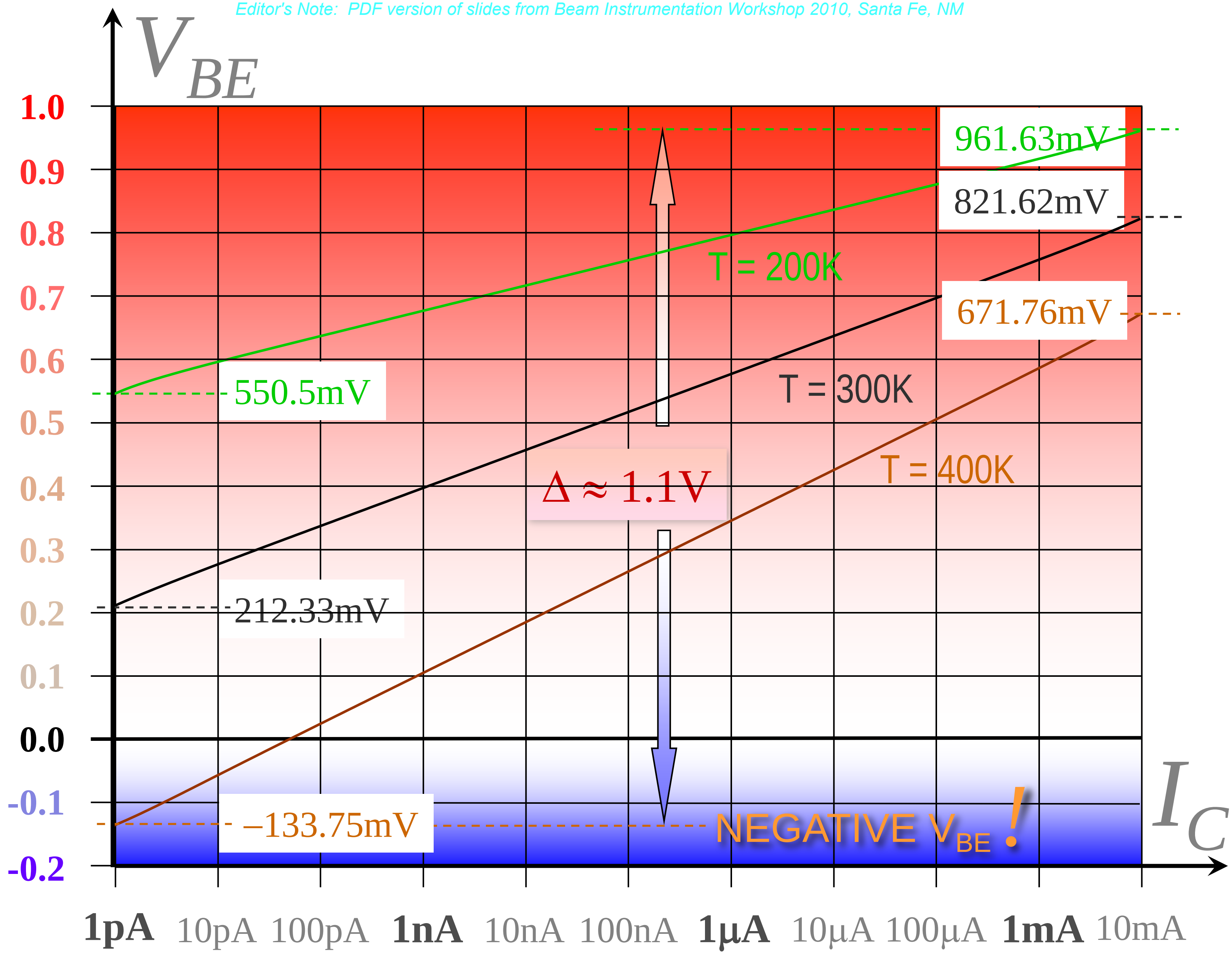


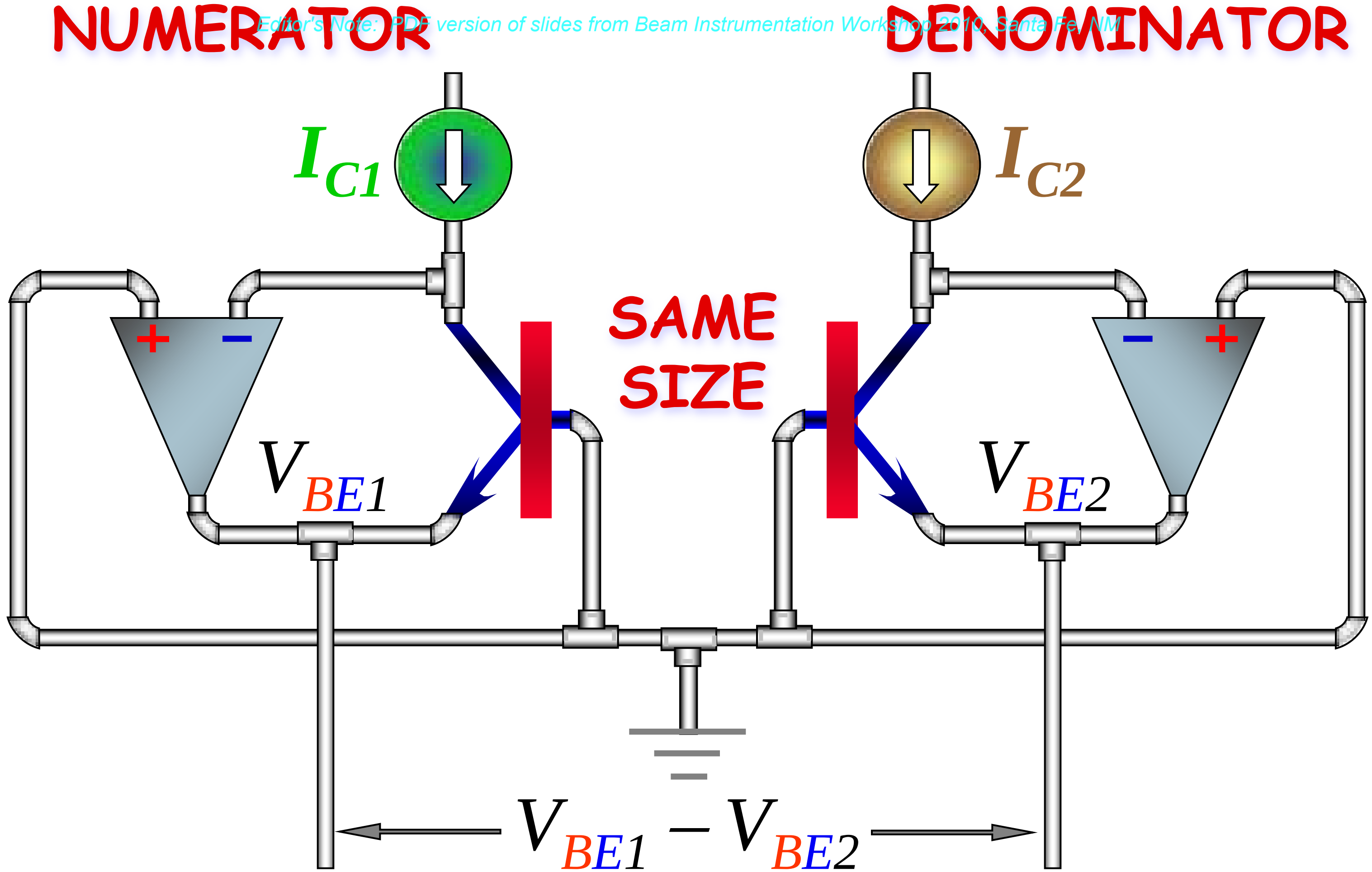


V_{BE}
1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0.0
-0.1
-0.2
961.63mV
821.62mV
T = 200K
671.76mV
550.5mV
T = 300K
T = 400K
Δ ≈ 1.1V
212.33mV
–133.75mV
NEGATIVE V_{BE} !
I_C
1pA
10pA
100pA
1nA
10nA
100nA
1µA
10µA
100µA
1mA
10mA

ELIMINATING THE VERTICAL SHIFT

The vertical shift in $V_{BE}(T)$ is due to the extreme change in $I_S(T)$ over temperature (the ratio can be as high as a trillion-to-one from -55°C to +125°C).

This shift is readily eliminated by using a second transistor, operating at a fixed reference current; this current and the size of this device determines the raw intercept of the logarithmic conversion.



$$V_{BE1} - V_{BE2} = V_K \log I_{C1} / I_S(T) - V_K \log I_{C2} / I_S(T)$$

$$= V_K \log I_{C1} / I_{C2}$$

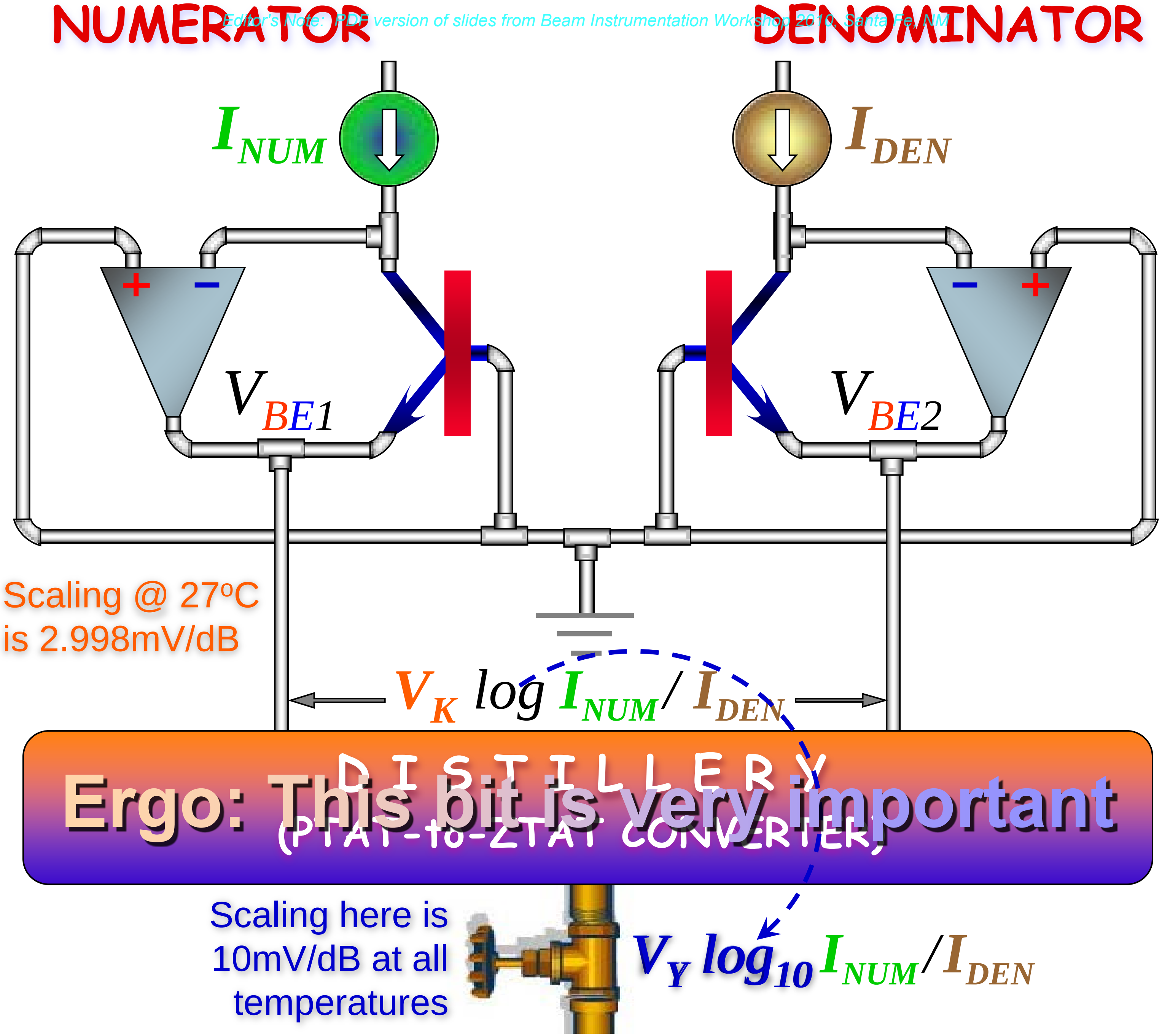
ELIMINATING THE CHANGE IN SLOPE

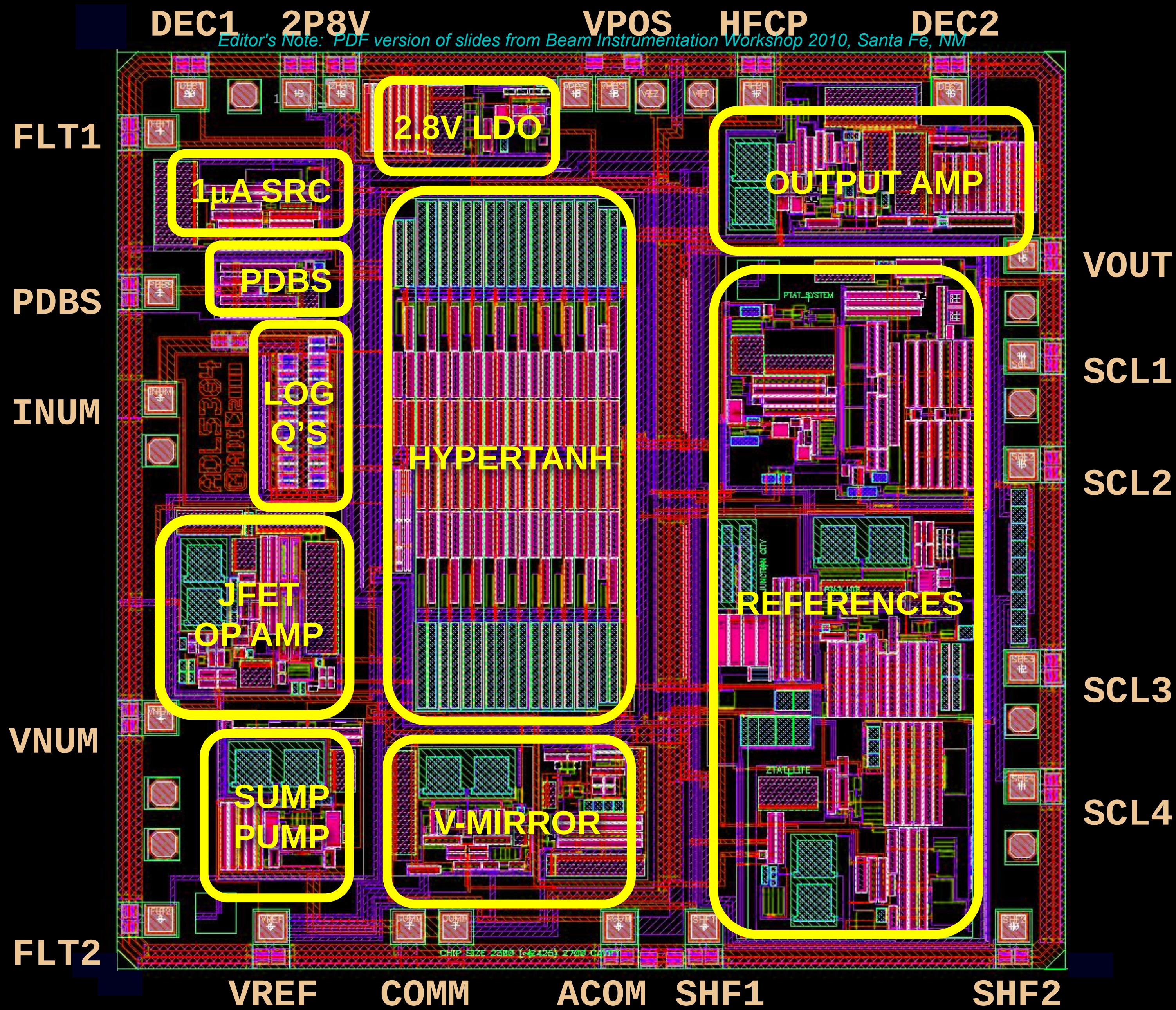
The variation in the slope of $V_{BE}(T)$ is due simply to the inherent PTAT scaling factor of $V_K = kT/q$:

$$\Delta V_{BE} = V_K \log I_{C1} / I_{C2}$$

We need to convert this factor from PTAT to ZTAT, that is, a slope that is invariant with temperature, re-evaluate it, now using the decibel basis of $\log_{10}$ and rename the two currents:

$$\mathbf{V_{LOG}} = \mathbf{V_Y} \log_{10} \mathbf{I_{NUM}} / \mathbf{I_{DEN}}$$


NUMERATOR
DENOMINATOR
I_{NUM}
I_{DEN}
V_{BE1}
V_{BE2}
Scaling @ 27°C is 2.998mV/dB
$V_K \log I_{NUM} / I_{DEN}$
DISTILLERY
(PTAT-to-ZTAT CONVERTER)
Ergo: This bit is very important
Scaling here is 10mV/dB at all temperatures
$V_Y \log_{10} I_{NUM}/I_{DEN}$



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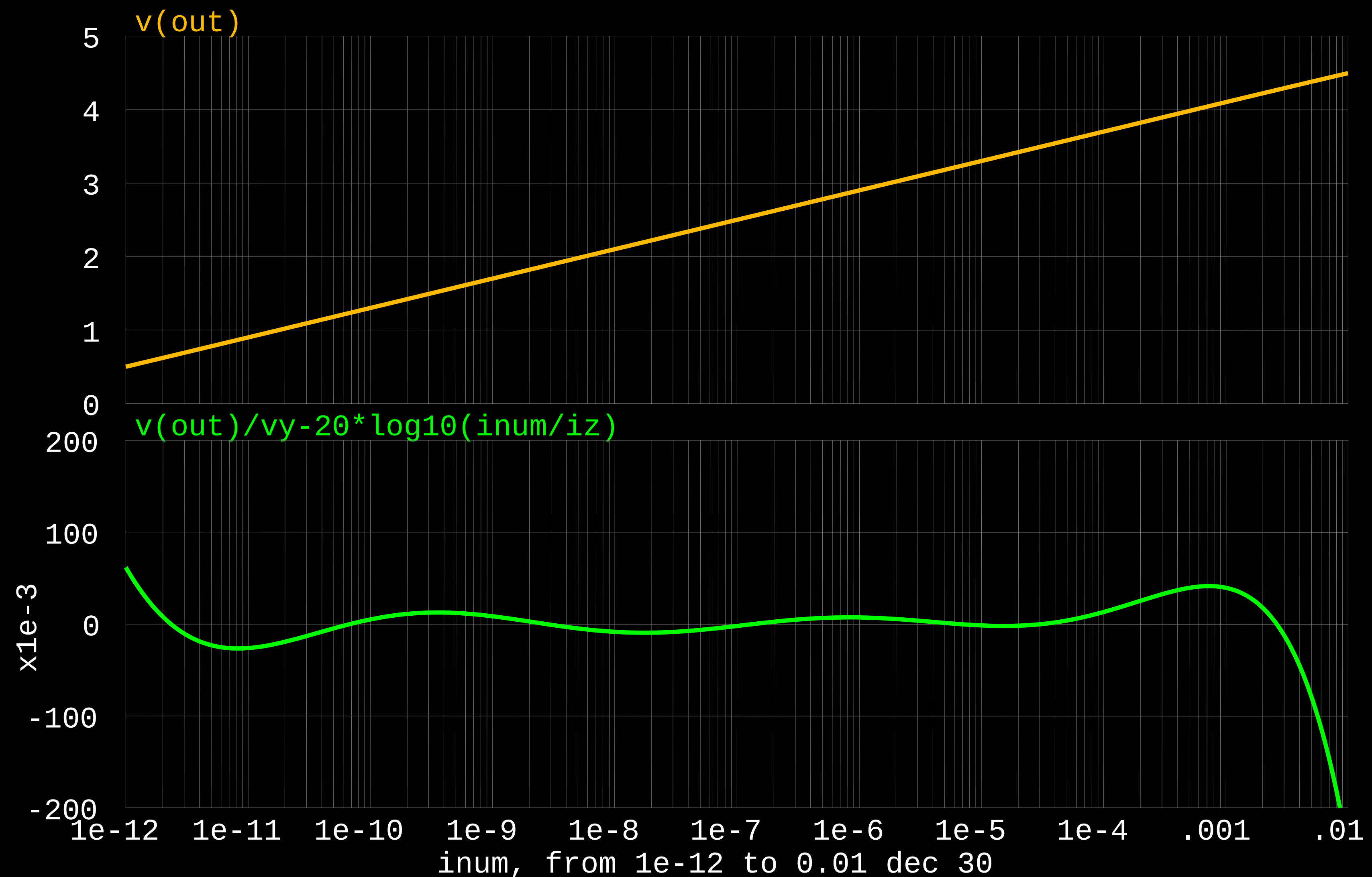
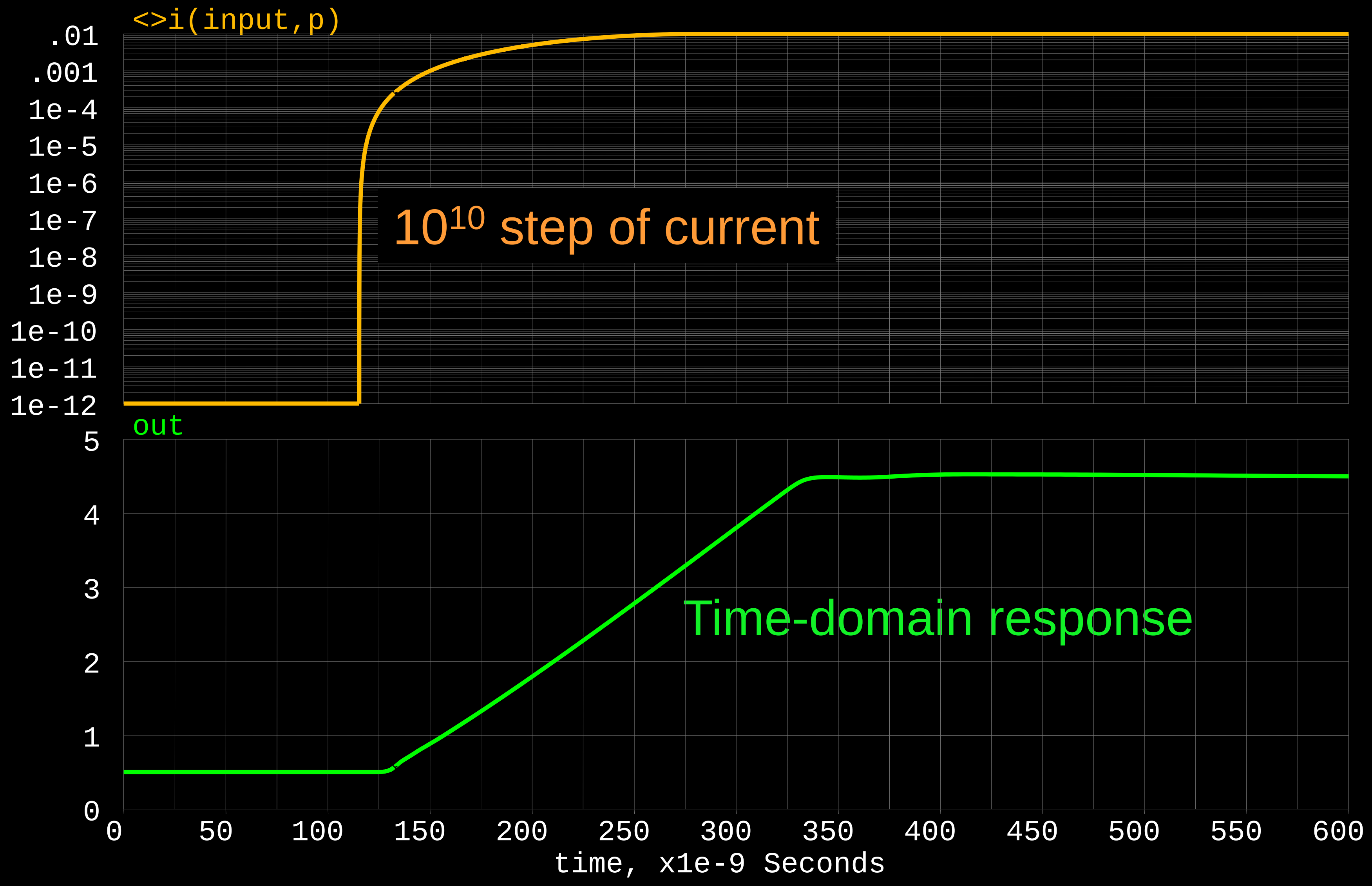
A 200-dB sweep of DC input current, from 1pA to 10mA, showing the scaling of 20mV/dB and very low log-conformance error (+/-0.05dB)

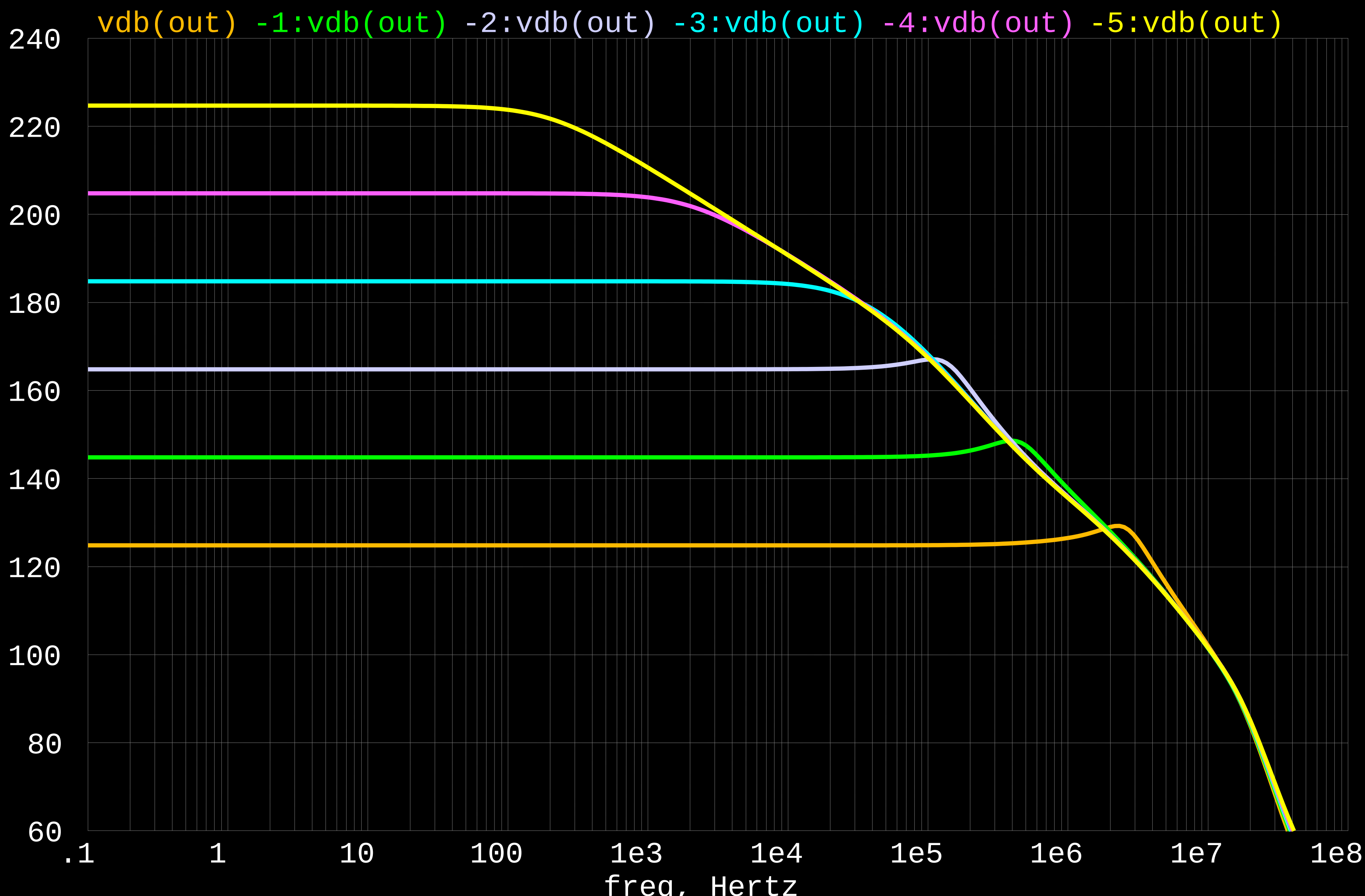


A 200-dB step of input current, from 1pA to 10mA, with a sine-squared rise-time of 100ns, is accomodated within a slewing time of 200ns



AC response for DC inputs of 1pA, 10pA, 100pA, 1nA, 10n and 100nA

Note the LF gain at 1pA is 225dB, corresponding to RT = 178 G-ohm



EXPONENTIAL AGC

- A VARIABLE-GAIN AMPLIFIER HAVING INVERSE EXPONENTIAL CONTROL OF GAIN (e^{-x}, that is, ‘Linear-in-dB’) …..
- …. DRIVES A DETECTOR CELL TO A FIXED SET-POINT AT ITS OUTPUT
- THIS SCHEME CAN OPERATE OVER A LARGE DYNAMIC RANGE (noise-limited)…
- … AND IS CAPABLE OF PROVIDING AN EXACT RMS (TRUE POWER) RESPONSE

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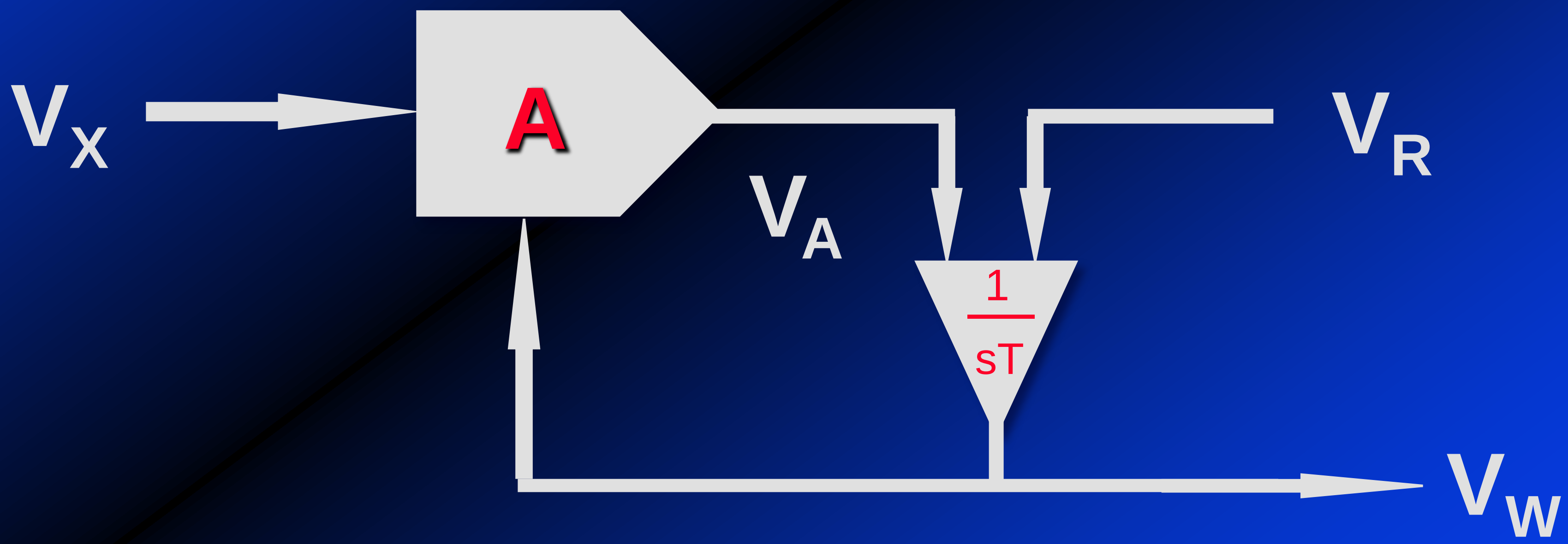
LOG-AMP BASED ON 'EXPONENTIAL AGC'

$$A = A_0 \exp(-V_W / V_Y)$$



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LOG-AMP BASED ON 'EXPONENTIAL AGC'

The amplifier output settles to equal V_R by the action of the loop: the mean input to the integrator must be forced to zero. Thus

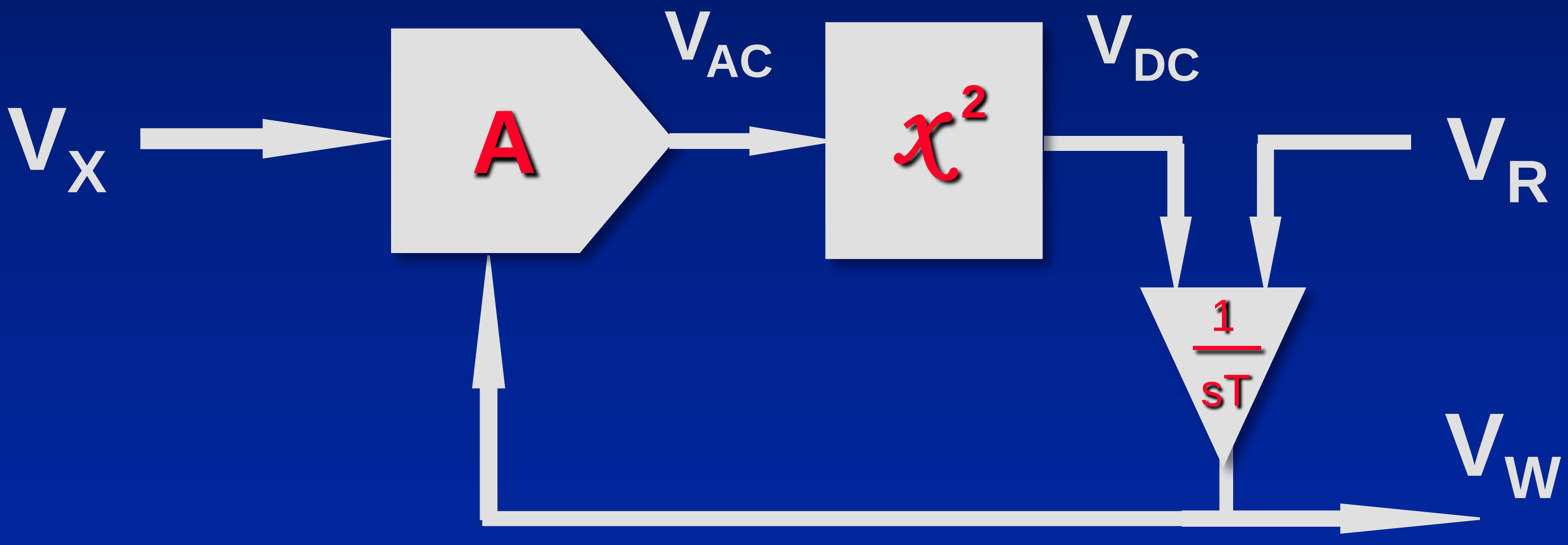
$$V_A = V_X A_0 \exp(-V_W/V_Y) \rightarrow V_R$$

Solving for V_W:

$$V_W = V_Y \log(V_X/V_Z) \qquad V_z = V_R/A_0$$

LOG-AMP BASED ON 'EXPONENTIAL AGC'

For AC signals, add a detector:



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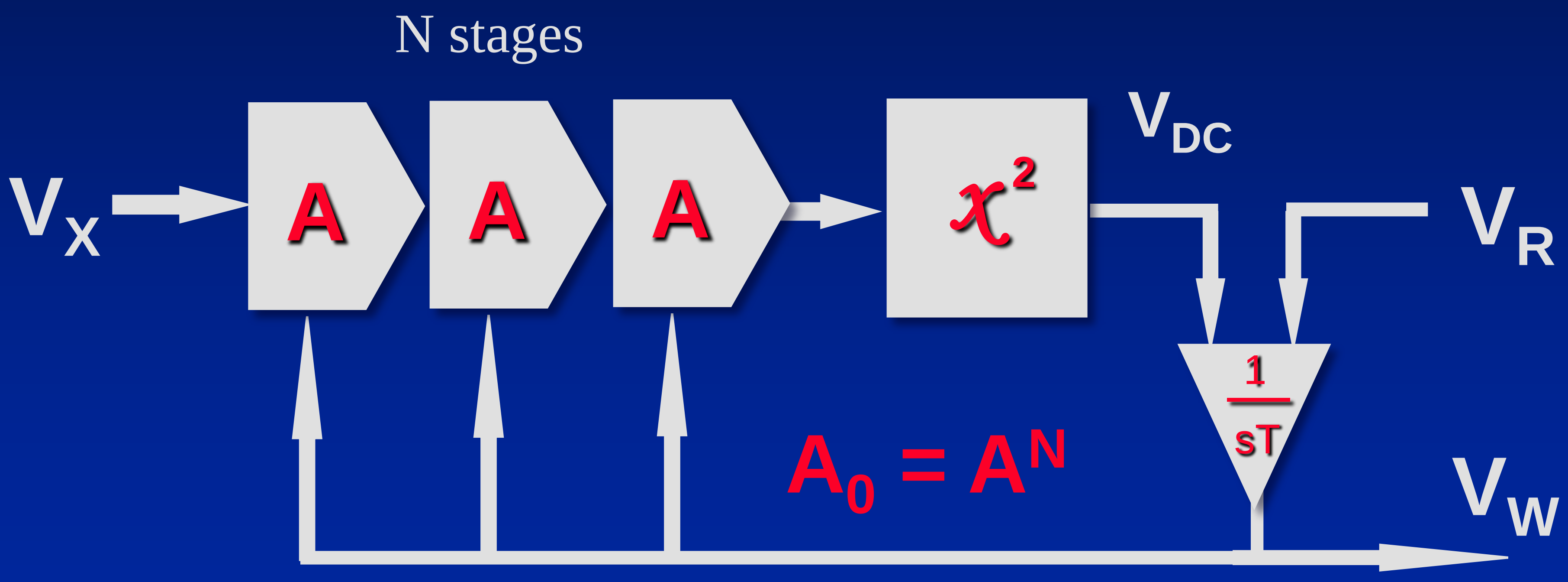
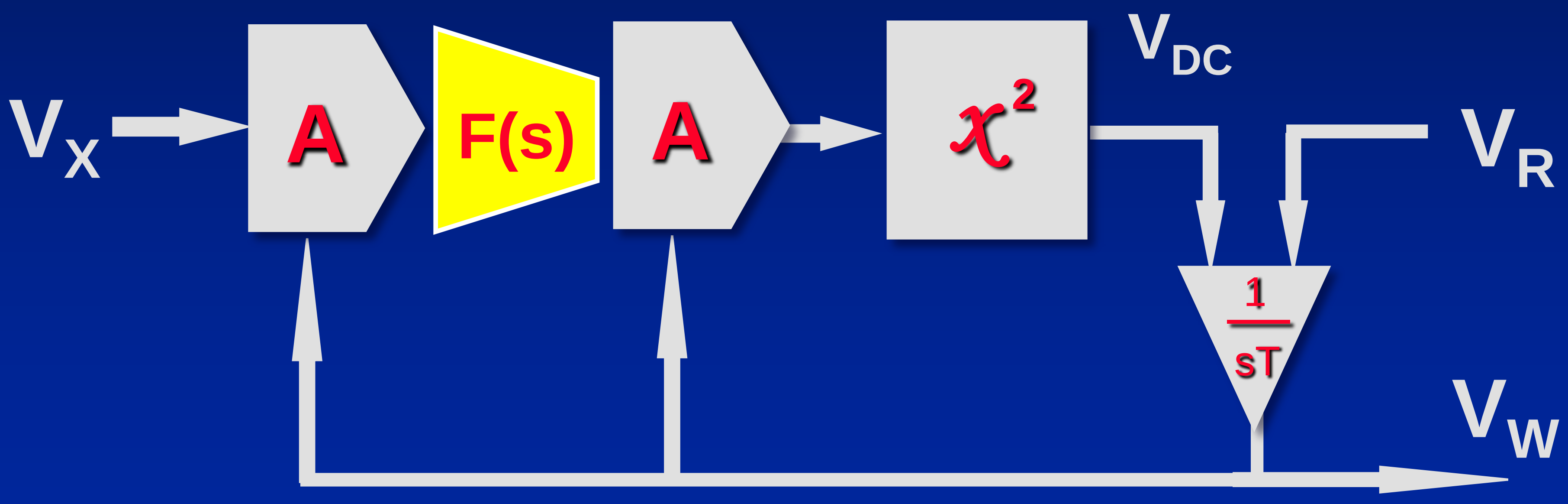
LOG-AMP BASED ON ‘EXPONENTIAL AGC’

For higher dynamic range and higher bandwidth, cascade several stages:



LOG-AMP BASED ON 'EXPONENTIAL AGC'

For even higher dynamic range, add filter(s):



120dB of range is readily obtainable

The X-AMP

- A PROPRIETARY VGA PRINCIPLE
- FUNDAMENTALLY "LINEAR-in-dB"
- USES FEEDBACK IN ORDER TO:
  ACCURATELY DETERMINE GAIN
  & MINIMIZE HF NONLINEARITIES
- GUARANTEES ULTRA-LOW NOISE
- EXHIBITS WIDE DYNAMIC RANGE
  FROM NOISE FLOOR (0.7μV RMS)
  TO TYPICALLY 1.4V RMS (106dB)

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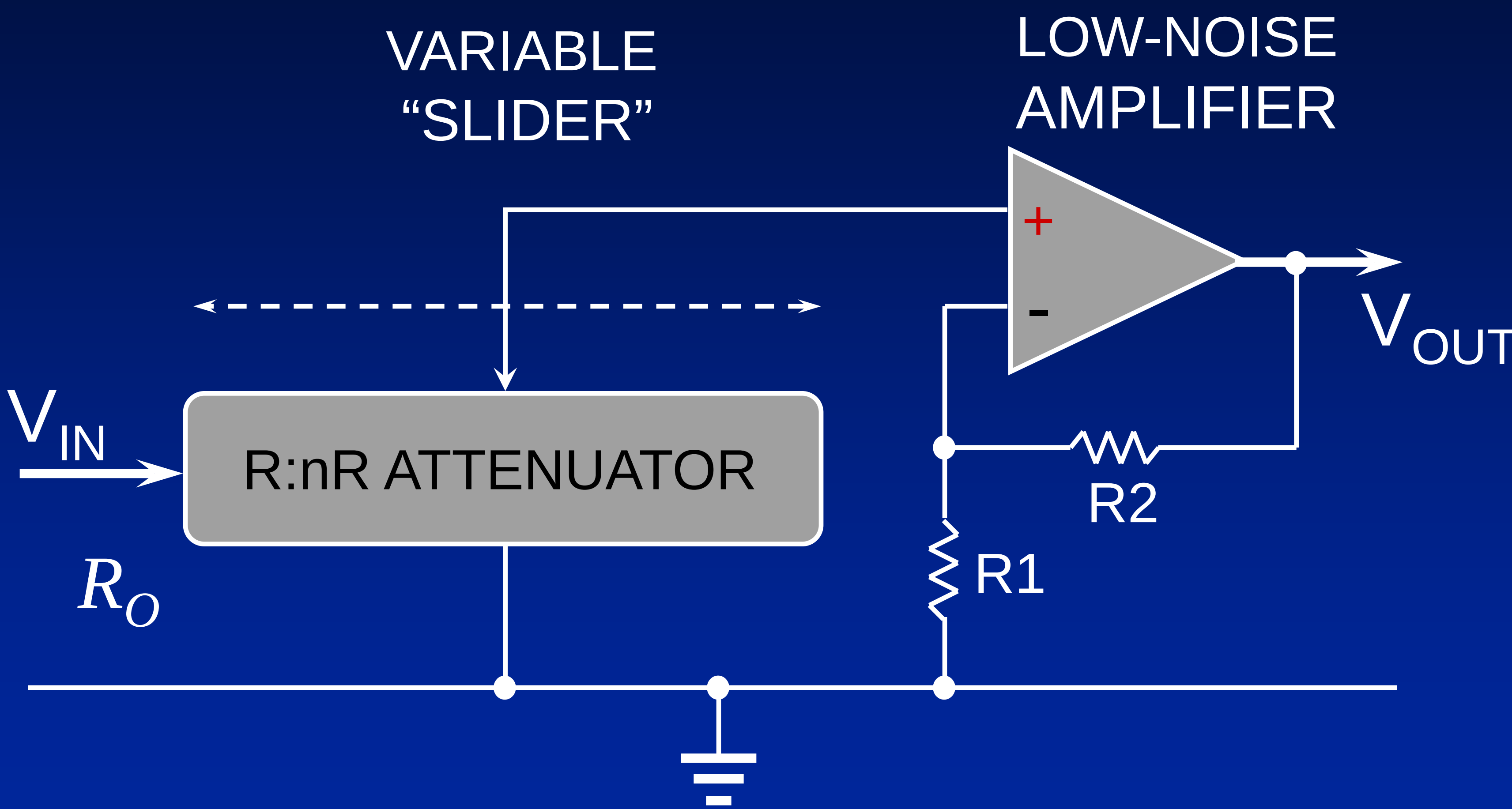
A BASIC X-AMP



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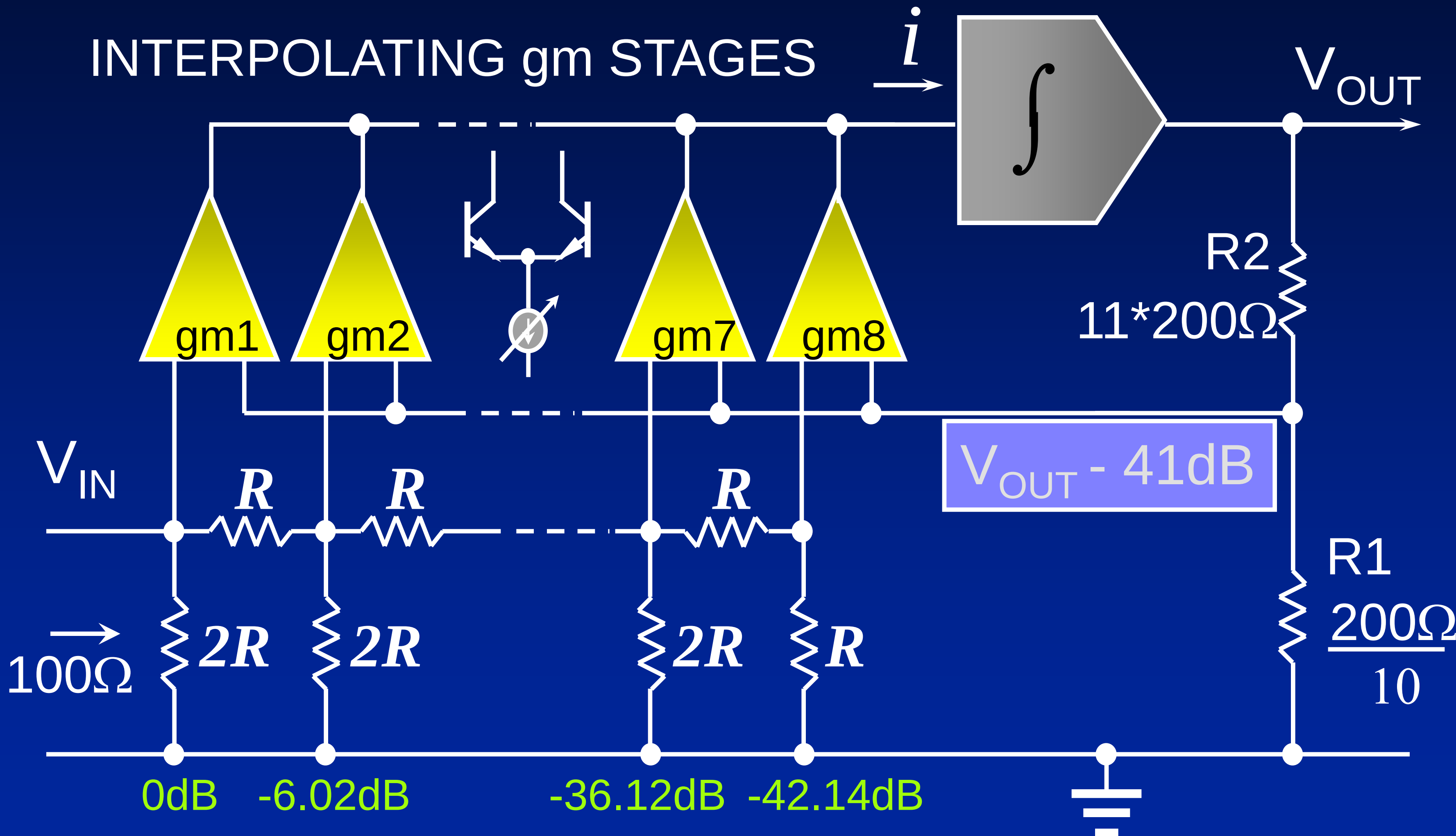
TYPICAL 8-STAGE X-AMP



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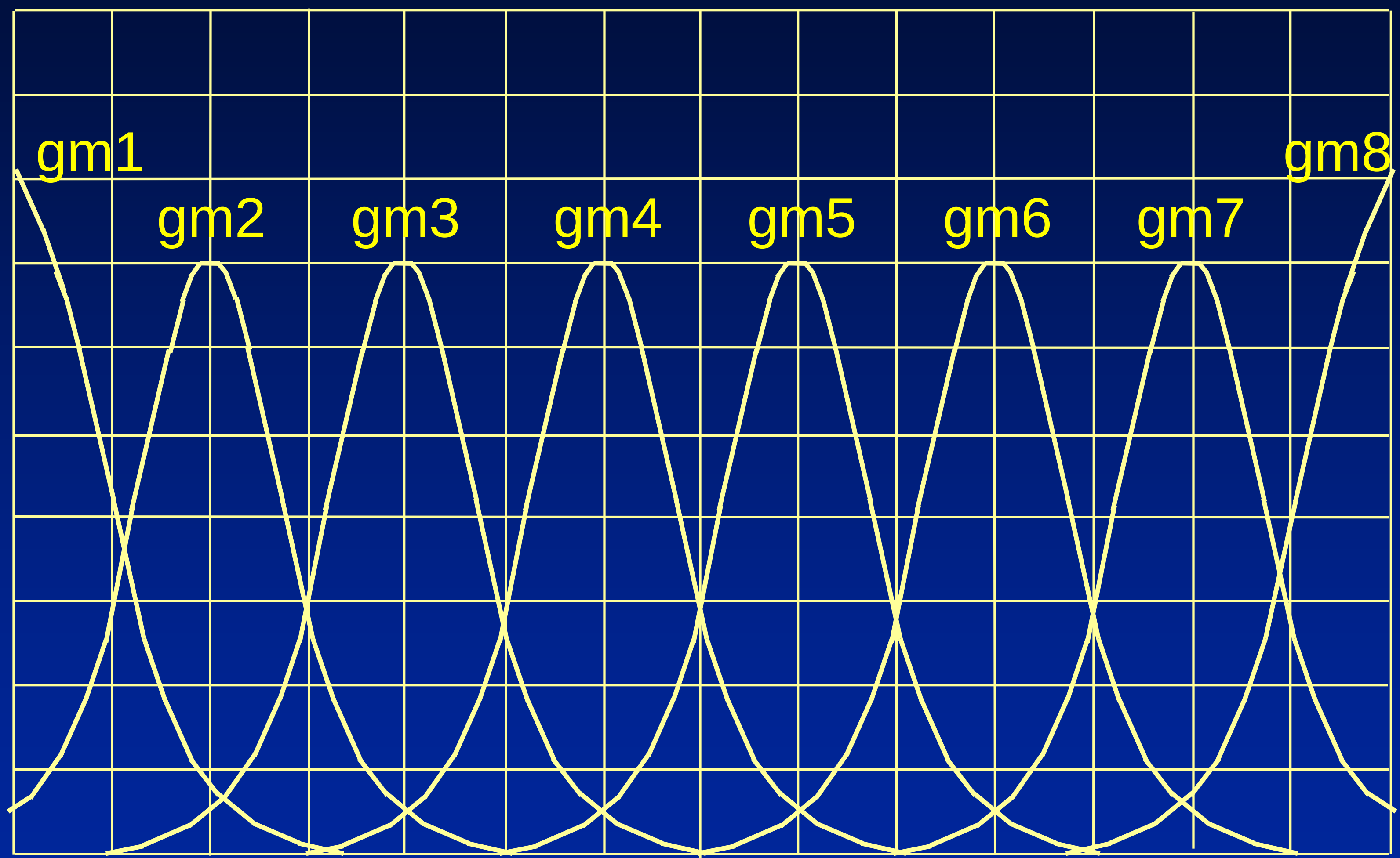
CURRENTS IN THE gm STAGES



INCREASING GAIN

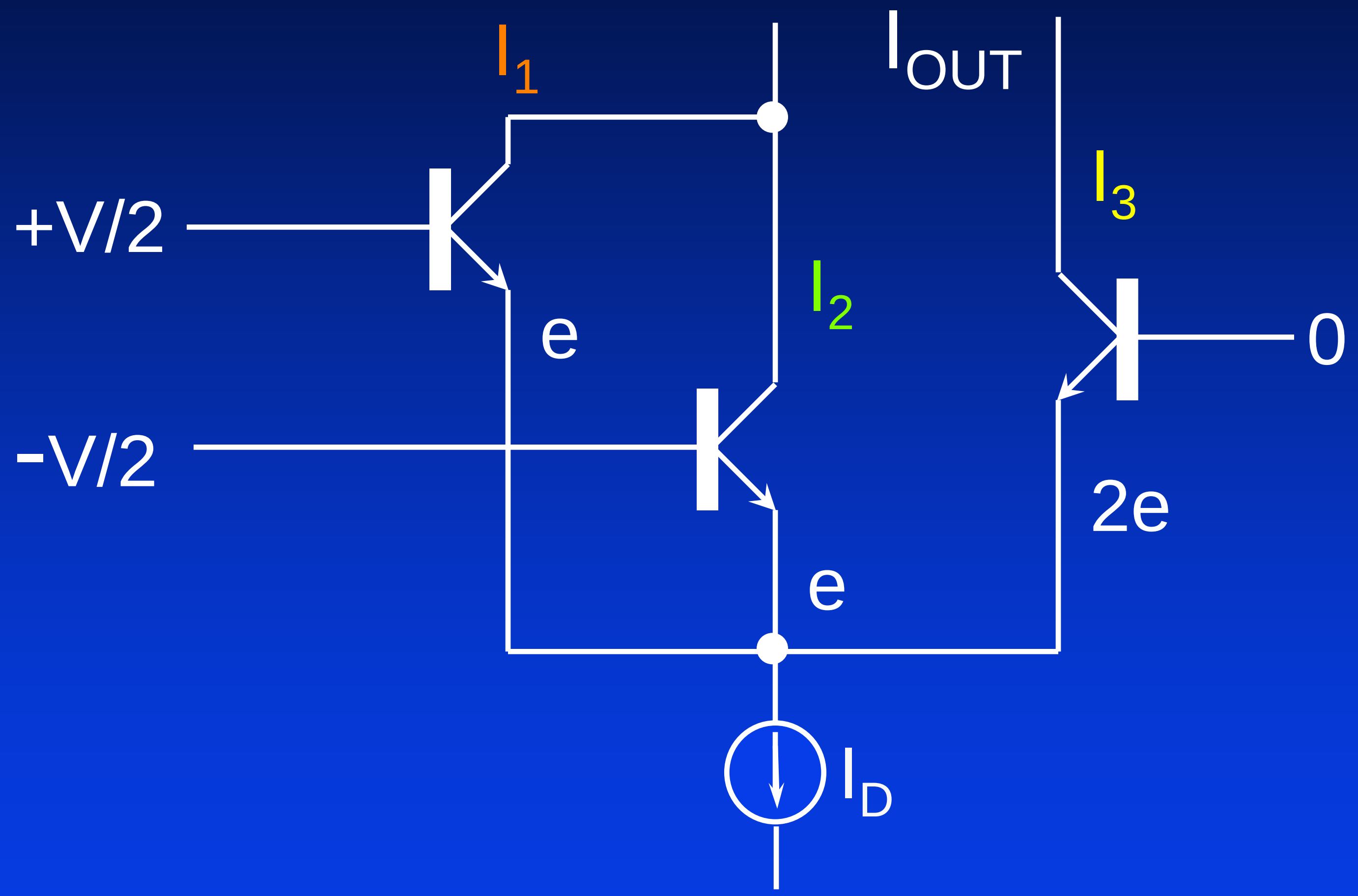
(MOVES ACTION TOWARDS FRONT)

AD8363

is an example of a log-amp that is also RMS-repsonding.

It can operate from a few Hz up to a specified 6 GHz, providing a 50-dB range and laser-trimmed calibration to absolute standards.

SQUARE-LAW DETECTOR



SQUARE-LAW DETECTOR

$$I_1 = I_D\, e^{u} / (e^{u} + e^{-u} + 2)$$

$$I_2 = I_D\, e^{-u} / (e^{u} + e^{-u} + 2)$$

$$I_3 = 2I_D / (e^{u} + e^{-u} + 2)$$

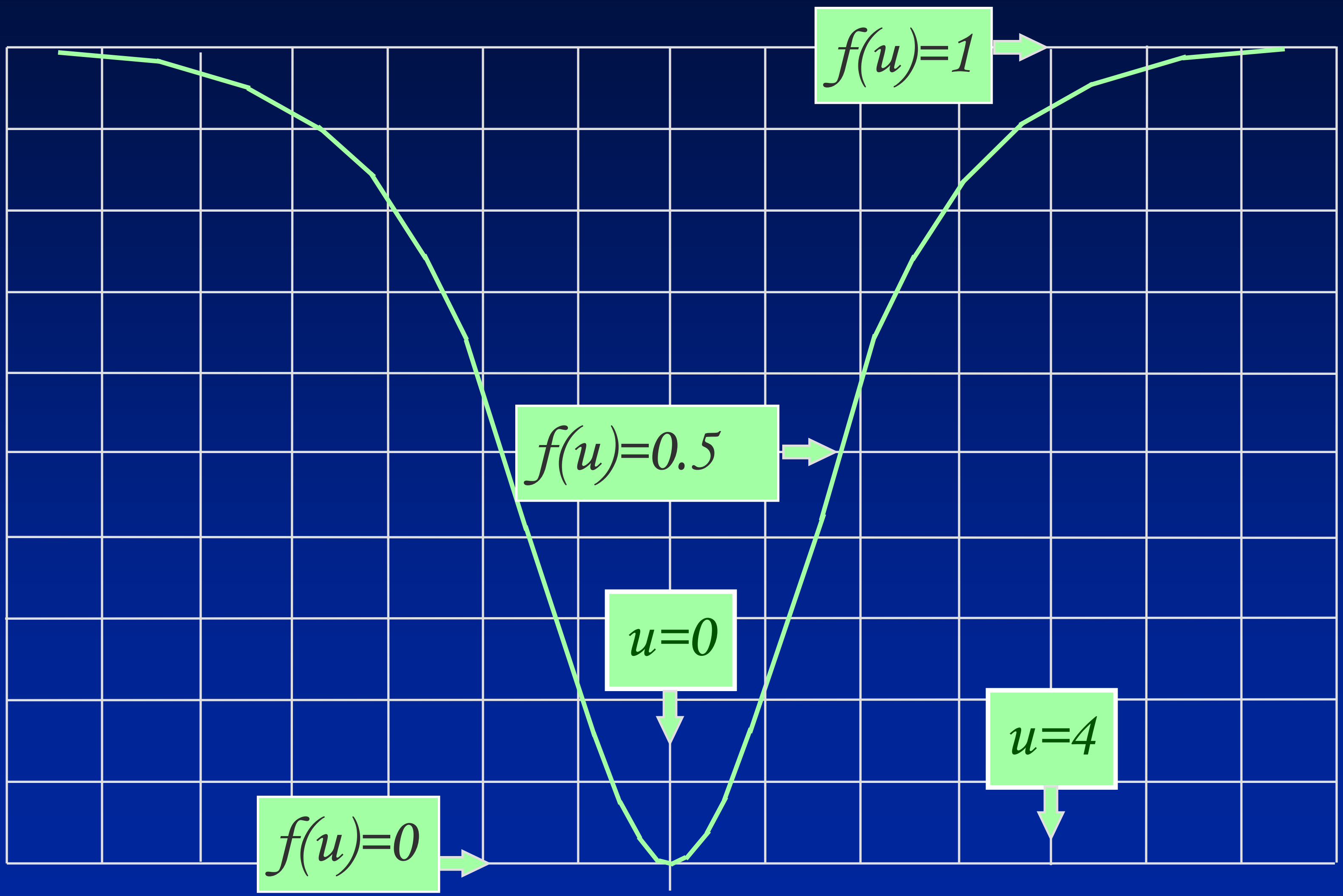
where

$u = V/2V_T$

$$I_{OUT} = I_1 + I_2 - I_3 = \frac{e^{u} + e^{-u} - 2}{e^{u} + e^{-u} + 2}\, I_D$$

Error is $< \pm 2.7\%$, $-2V_T < V < +2V_T$

$$f(u) = \frac{e^{u} + e^{-u} - 2}{e^{u} + e^{-u} + 2} = 1 - \mathrm{sech}^2(u/2)$$



END OF OVERVIEW

PROGRESSIVE COMPRESSION

- A MAJOR CLASS, with many sub-types
- THE METHOD USED FOR PRACTICALLY ALL WIDEBAND LOG-AMPS, UP TO SHF
- THE BACKBONE IS A CHAIN OF SIMPLE AMPLIFIER CELLS
- FUNCTION IS A TYPE OF PIECEWISE LINEAR APPROXIMATION, but the “law conformance error” may be as low as 0.1dB

PROGRESSIVE COMPRESSION

THE LITERATURE ON LOG AMP DESIGN IS SCANT, EMPIRICAL, AND SURPRISINGLY QUIET ON THE CRUCIAL MATTER OF SCALING.

THE ANALYSES NOW PRESENTED FLOW DIRECTLY FROM A SIMPLE, FUNDAMENTAL STARTING POINT. NO GUESSING NEEDED*!*

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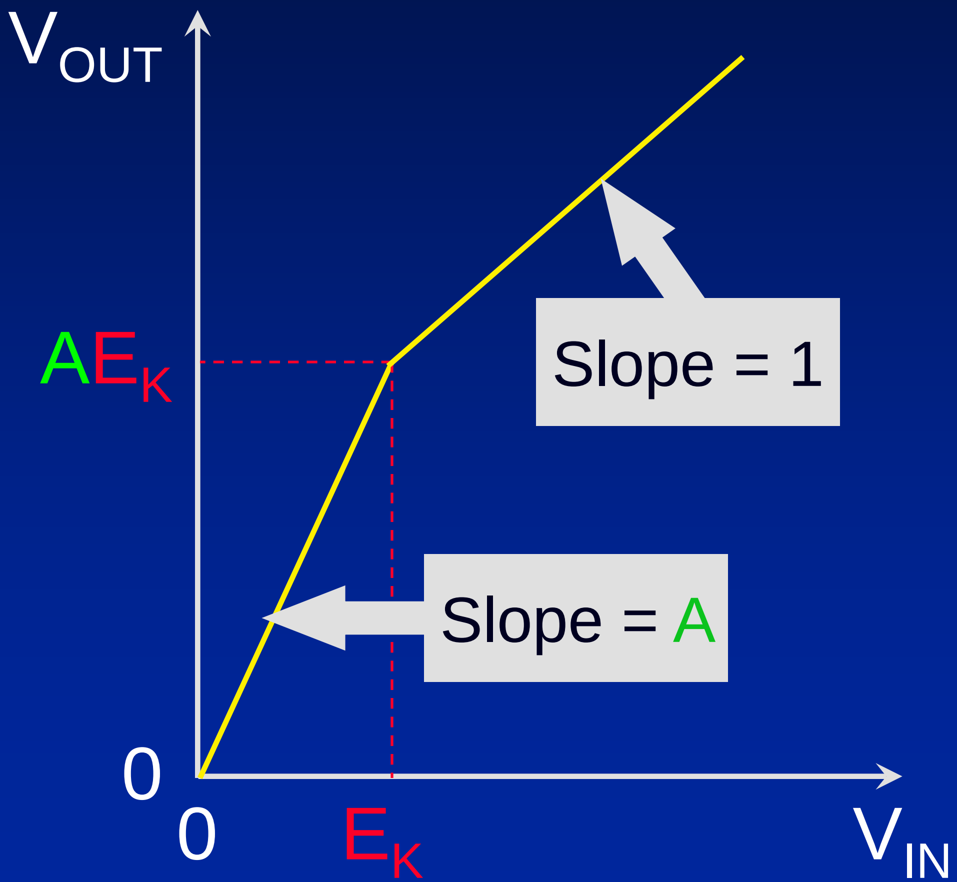
THE A/1 AMPLIFIER

SYMBOL



THE A/1 AMPLIFIER

$$V_{OUT} = AV_{IN}$$

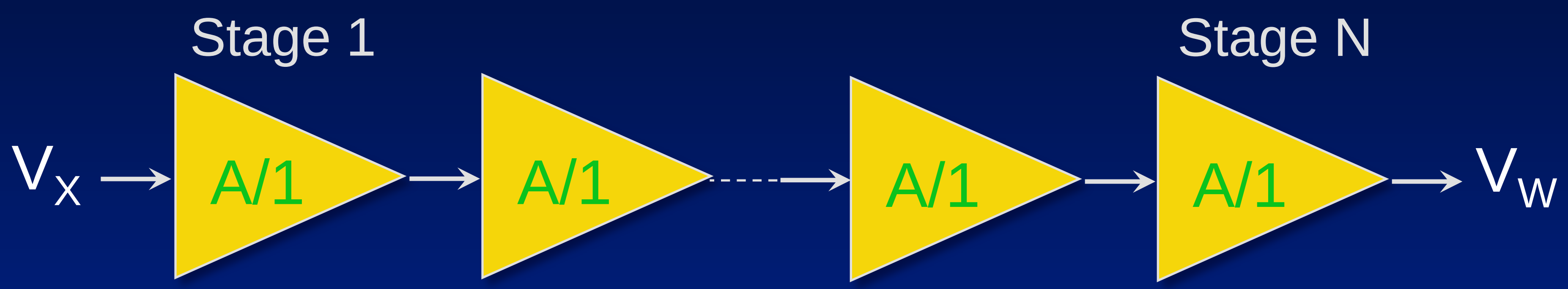
for $V_{IN} < E_K$

$$V_{OUT} = (A-1)\,E_K + V_{IN}$$

for $V_{IN} > E_K$

Note:
The amplifier function is usually odd-order symmetric

A CASCADE OF N, A/1 CELLS



- WE'LL IMPLEMENT A BASEBAND LOG-AMP
- USING PIECEWISE-LINEAR APPROXIMATION
- CASCADE HAS VERY HIGH OVERALL GAIN (A^N)

A CASCADE OF N, A/1 CELLS

THE FUNCTION WILL BE

$$V_W = V_Y \log \frac{V_X}{V_Z}$$

but.....

WHERE DO V_Y & V_Z “COME FROM”?

A CASCADE OF N, A/1 CELLS

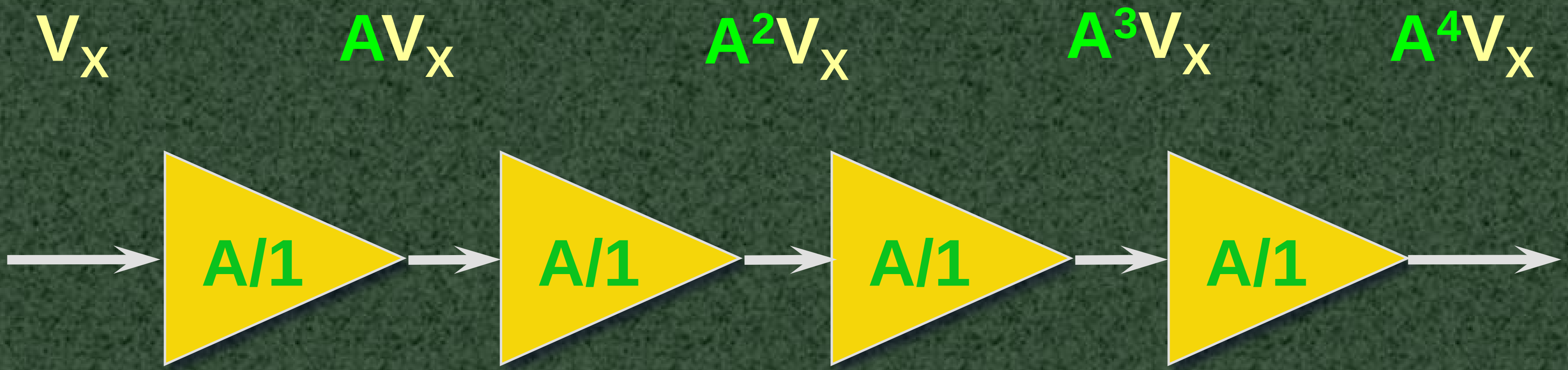
WE MUST CONCLUDE THAT

$$V_Y = y E_K$$

$$V_Z = z E_K$$

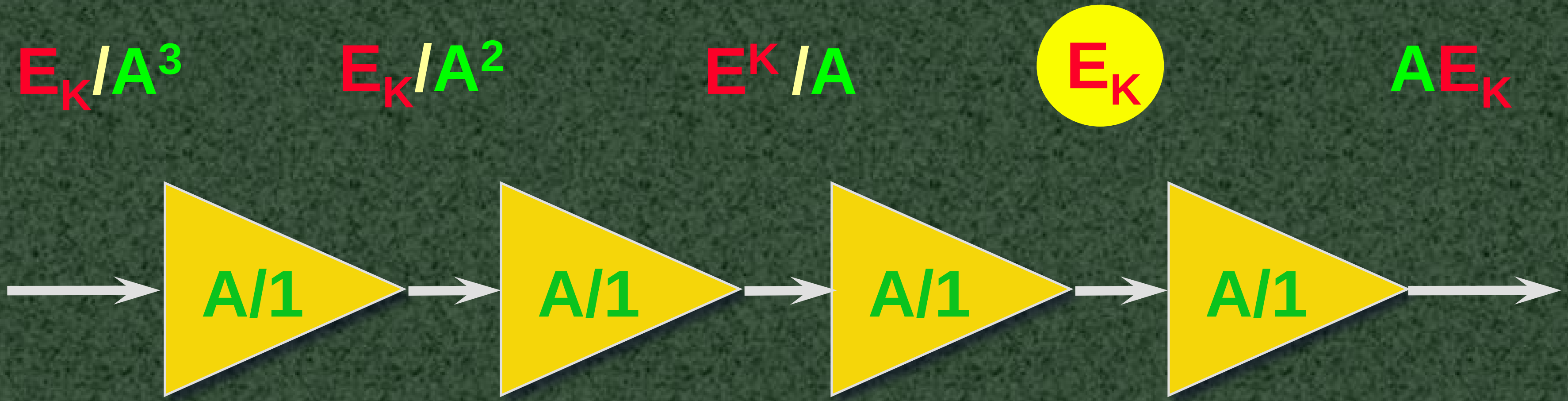
where y and z must be linear functions of A and N

FOUR-STAGE EXAMPLE



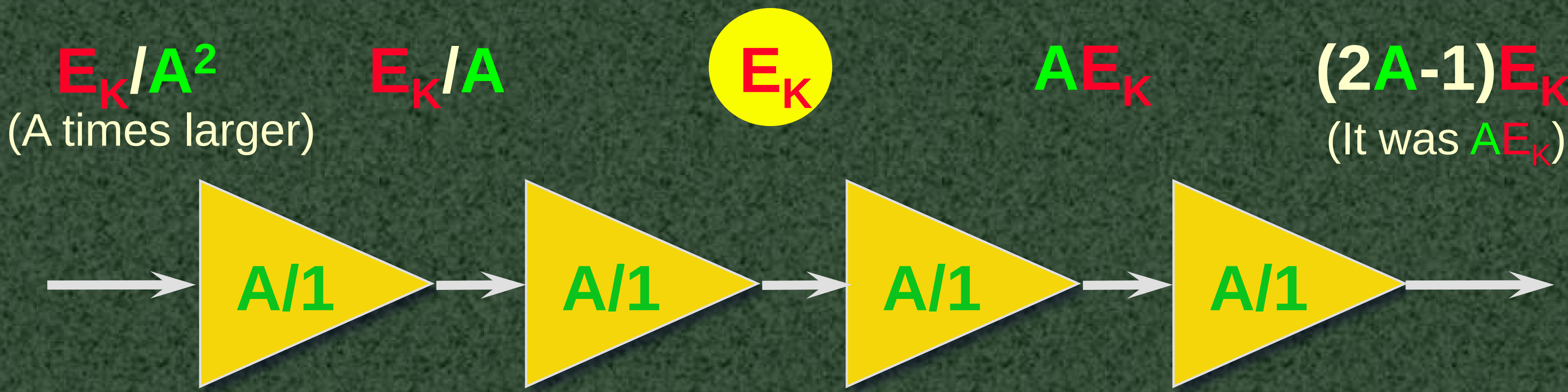
- INPUT VOLTAGE VERY SMALL
- ALL STAGES REMAIN LINEAR
- INTERNAL VOLTAGE E_K IS HIDING

FOUR-STAGE EXAMPLE



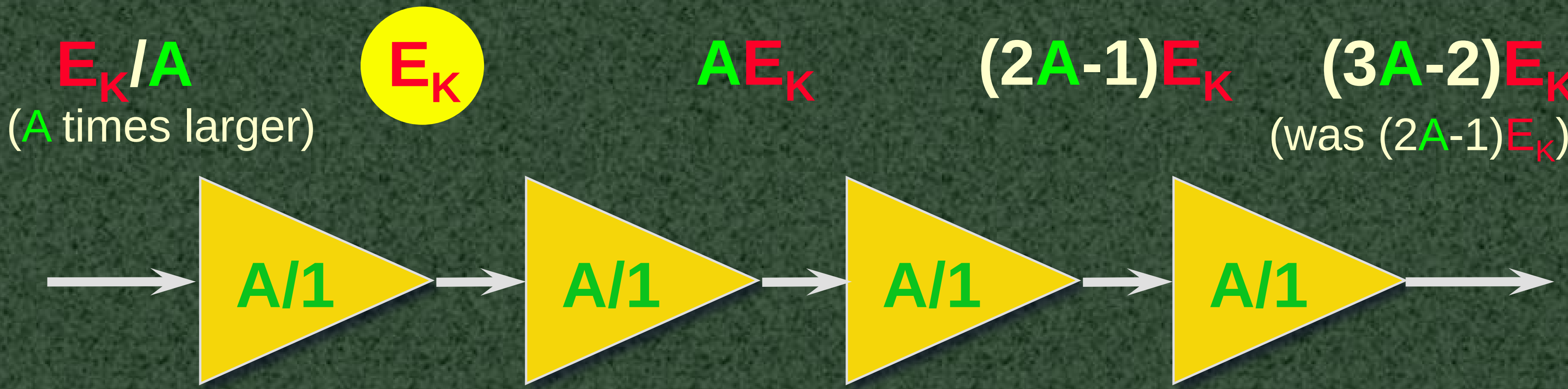
- FIRST TRANSITION POINT REACHED
- BOTH OUTPUT AND INPUT DEFINED
- VOLTAGE E_K NOW IN EVIDENCE

FOUR-STAGE EXAMPLE



- SECOND TRANSITION POINT
- OUTPUT HAS INCREASED BY $(A-1)E_K$
- INPUT INCREASED BY THE <u>RATIO</u> A

FOUR-STAGE EXAMPLE



- THIRD TRANSITION POINT
- OUTPUT HAS INCREASED BY $(A-1)E_K$
- INPUT INCREASED BY THE RATIO A

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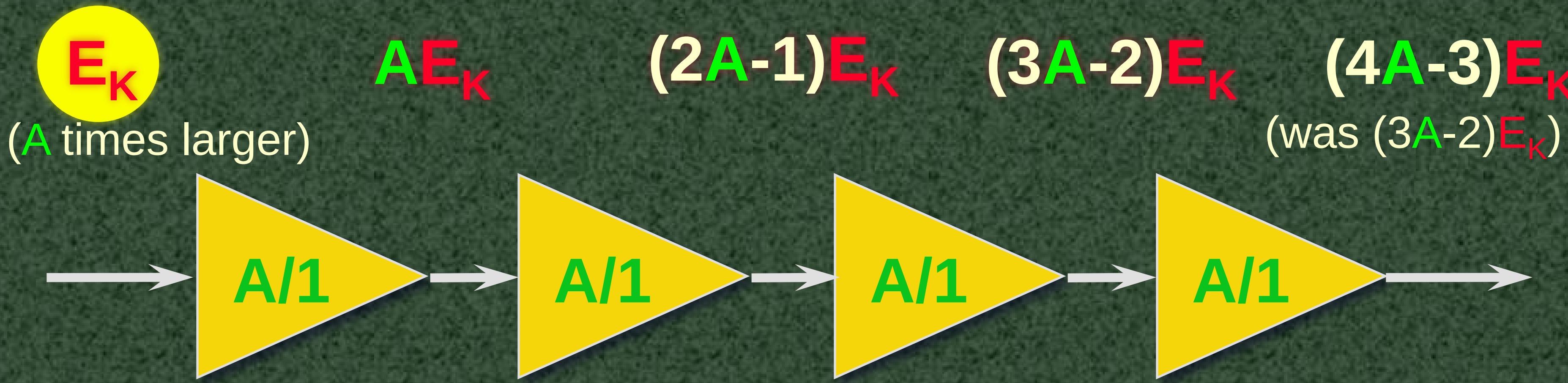
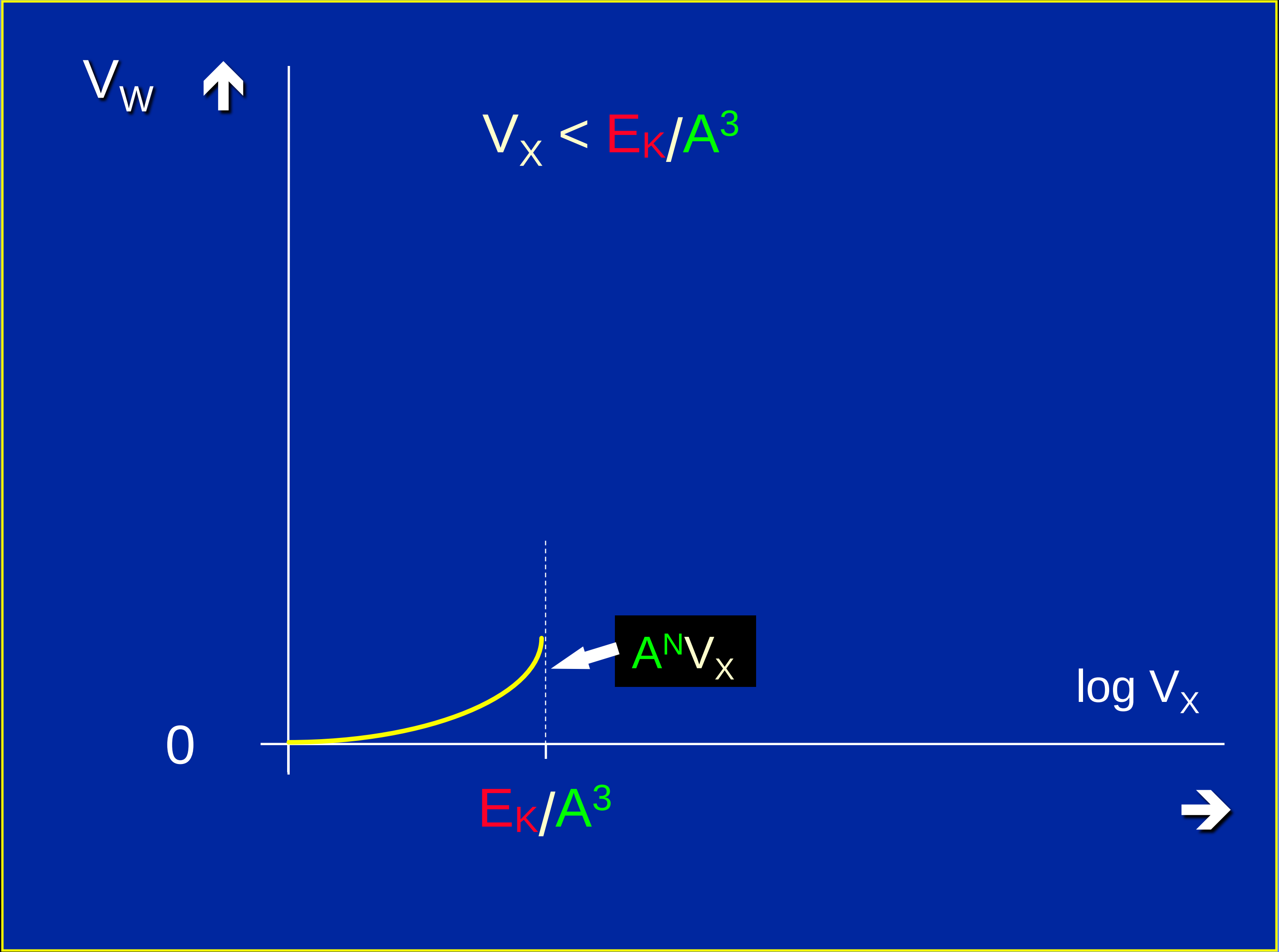
FOUR-STAGE EXAMPLE

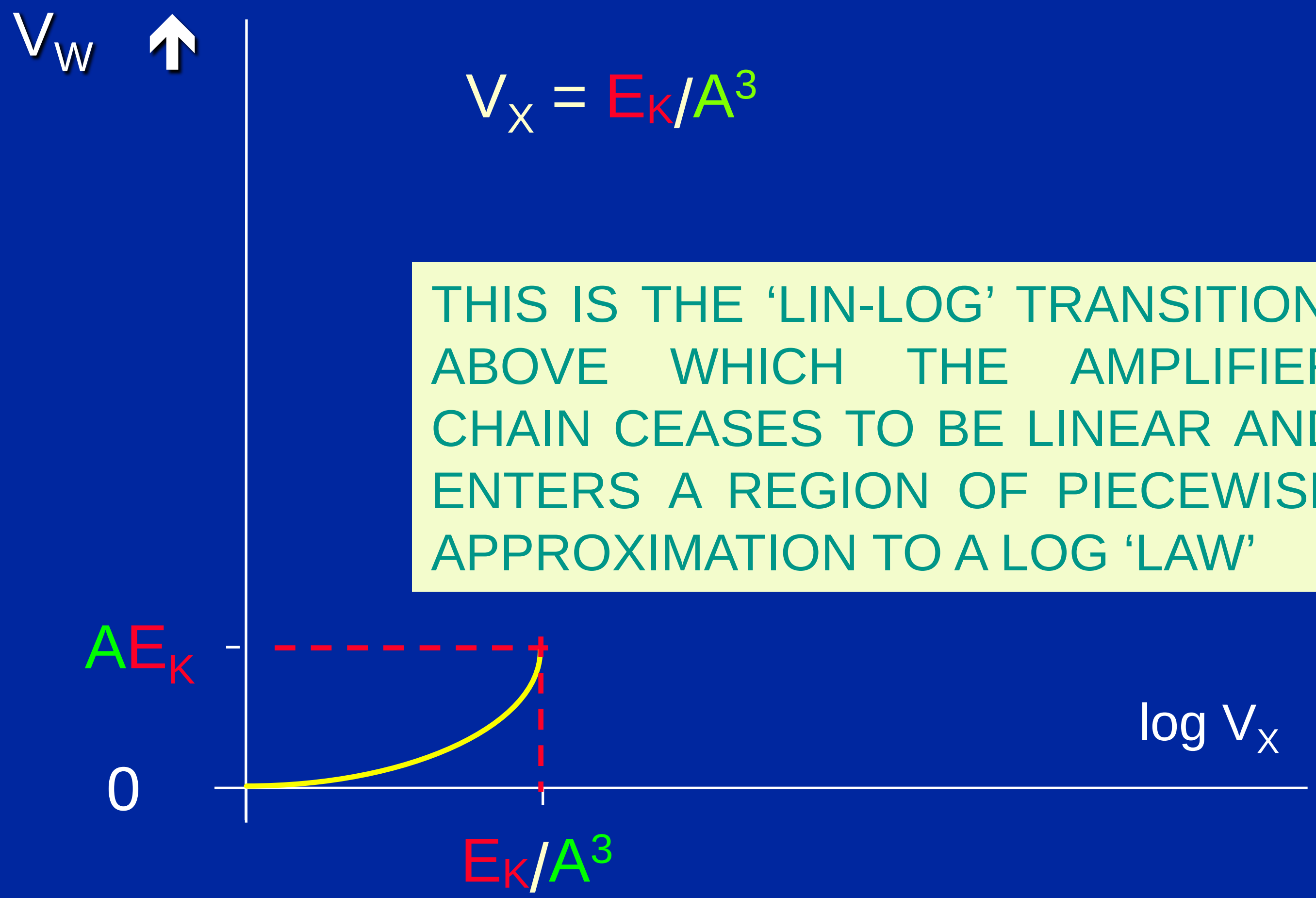


- FOURTH & FINAL TRANSITION POINT
- OUTPUT HAS INCREASED BY $(A-1)E_K$
- INPUT HAS INCREASED BY <u>RATIO</u> A

BUILDING THE LOGARITHMIC FUNCTION



BUILDING THE LOGARITHMIC FUNCTION



THIS IS THE 'LIN-LOG' TRANSITION. ABOVE WHICH THE AMPLIFIER CHAIN CEASES TO BE LINEAR AND ENTERS A REGION OF PIECEWISE APPROXIMATION TO A LOG 'LAW'

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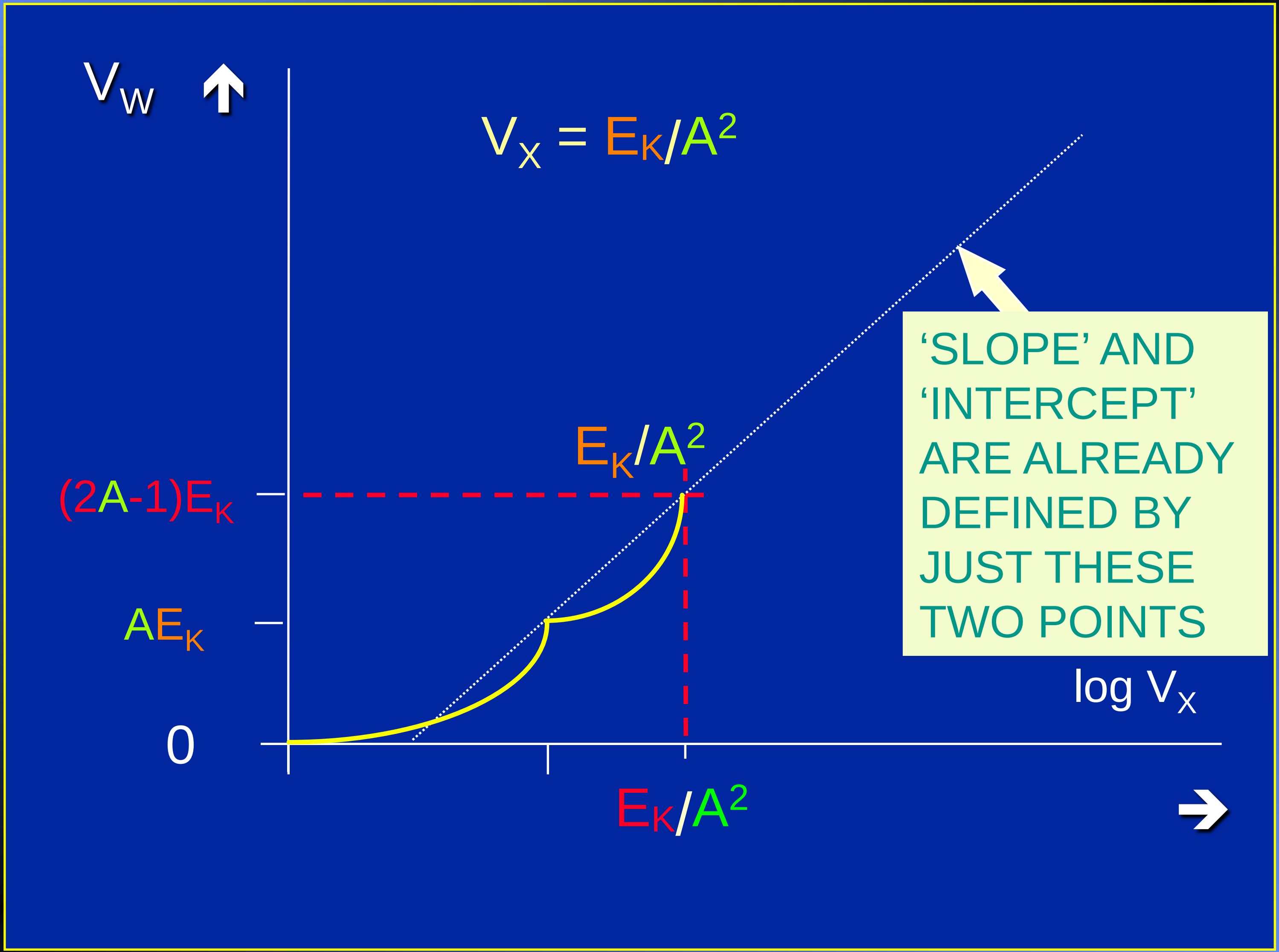
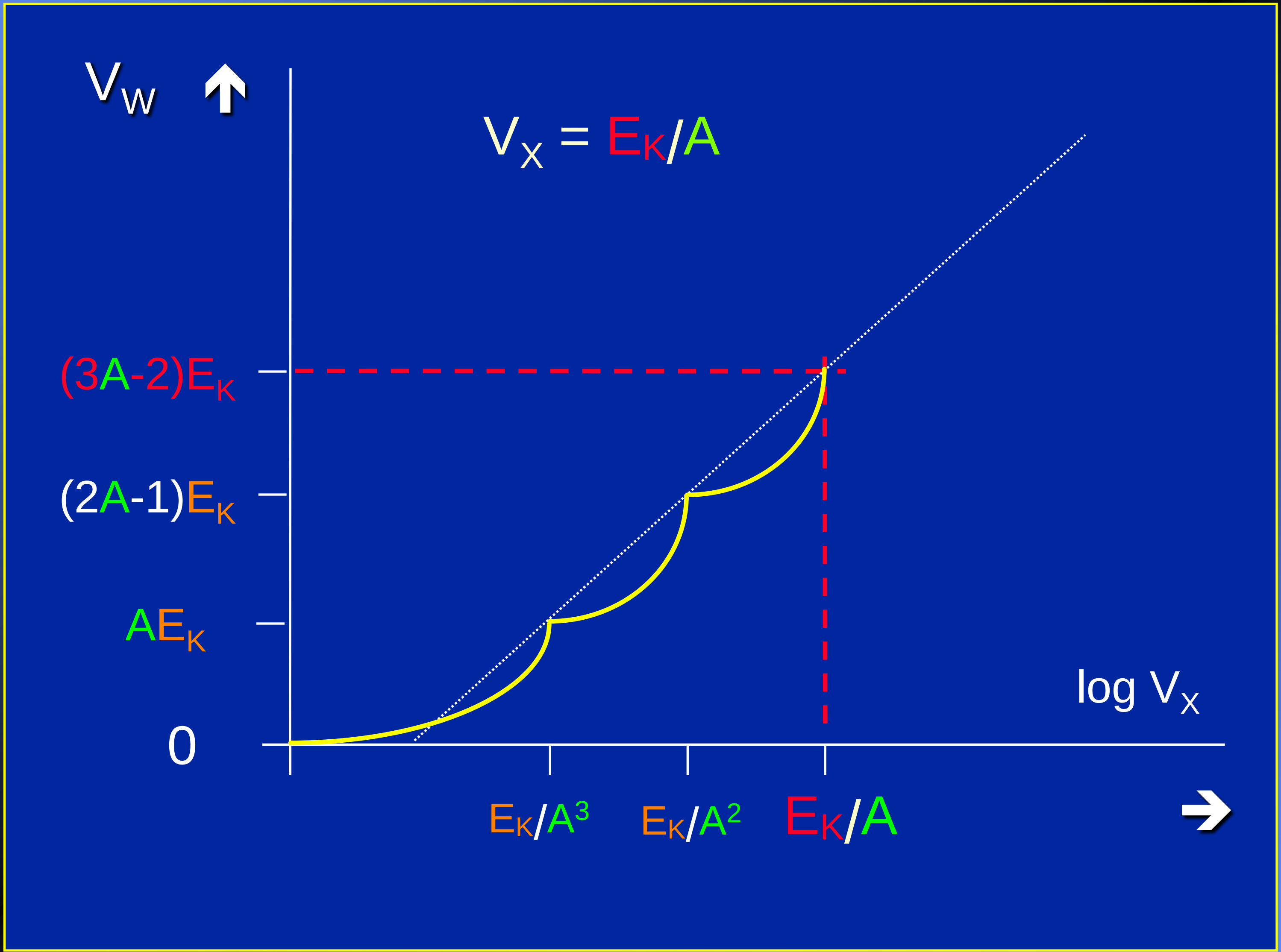
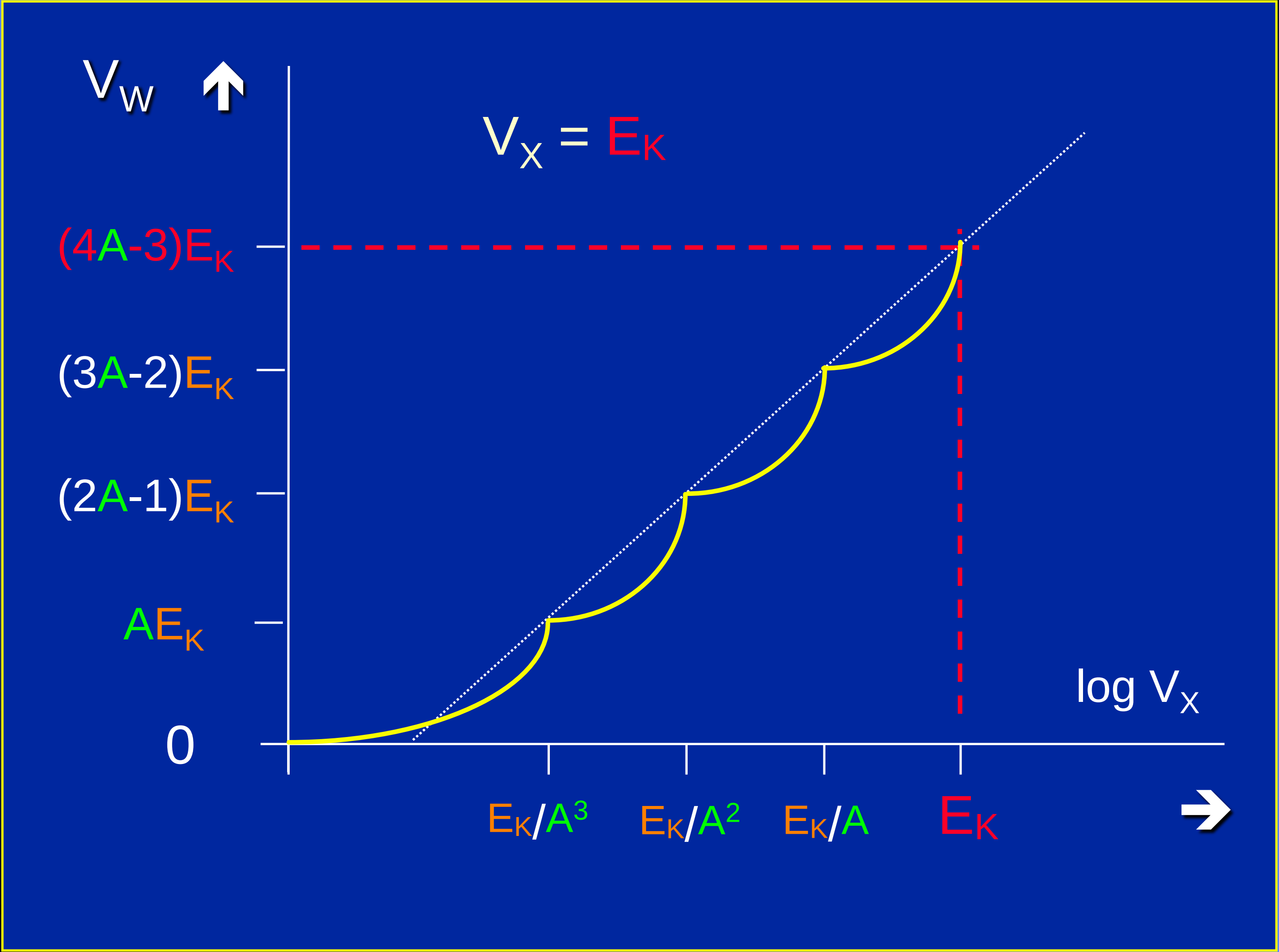
BUILDING THE LOGARITHMIC FUNCTION



BUILDING THE LOGARITHMIC FUNCTION



BUILDING THE LOGARITHMIC FUNCTION



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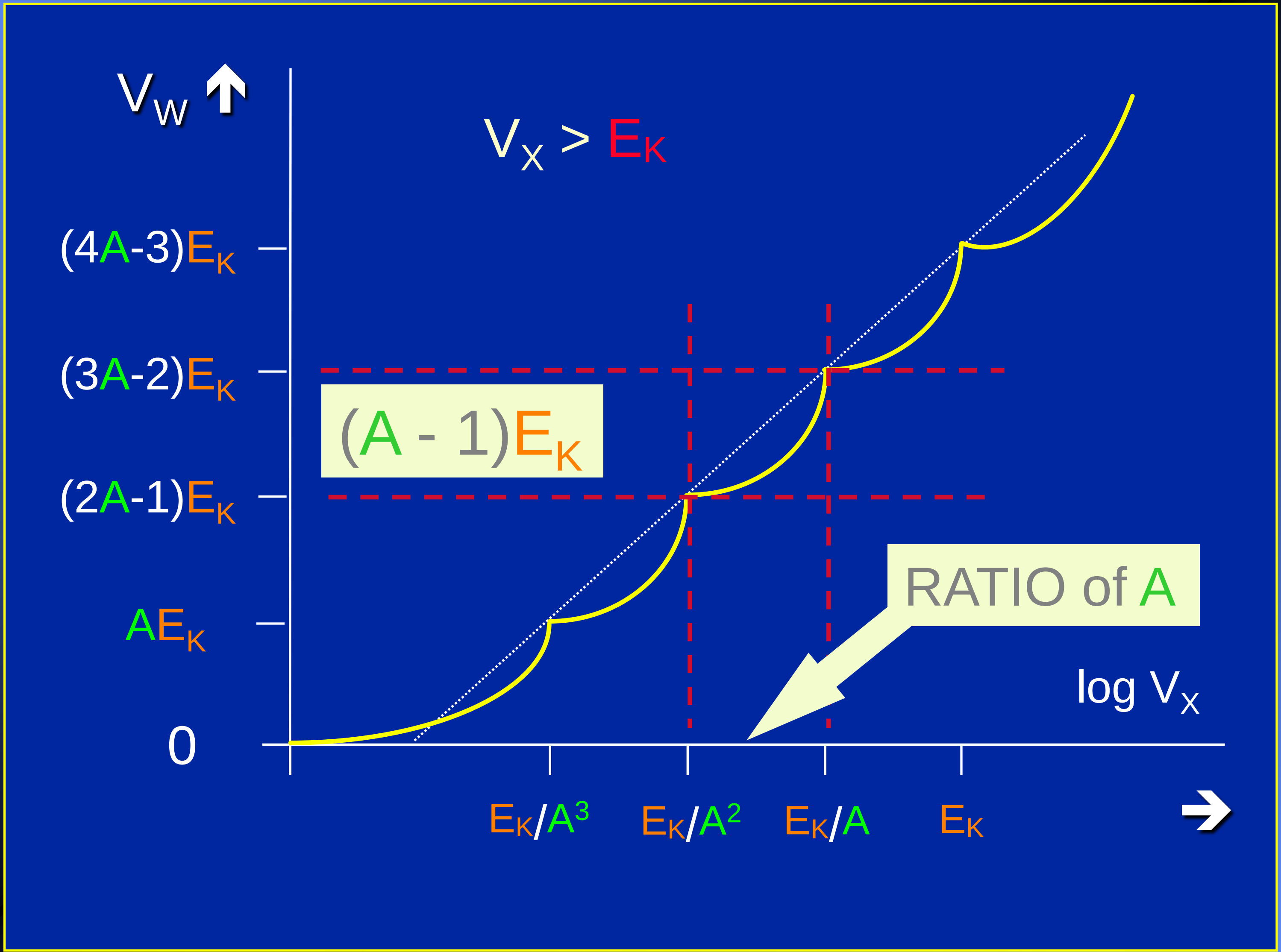
BUILDING THE LOGARITHMIC FUNCTION



SLOPE CALCULATION

THE OUTPUT CHANGES BY $(2A-1)E_K$ WHEN THE INPUT CHANGES BY THE *RATIO* A

$$V_Y = \frac{(2A-1)E_K}{\lg t\,(A)} \quad \text{(volts/decade)}$$

INTERCEPT CALCULATION

$$V_W = V_Y \,\mathrm{lgt}\, \frac{V_X}{V_Z}$$

SOLVE FOR V_Z BY SUBSTITUTING

$$V_Y = \frac{(2A-1)E_K}{\mathrm{lgt}\,(A)}, \quad V_X = E_K/(N-1), \quad V_W = AE_K$$

$$V_Z = \frac{E_K}{A^{N + 1/(A-1)}}$$

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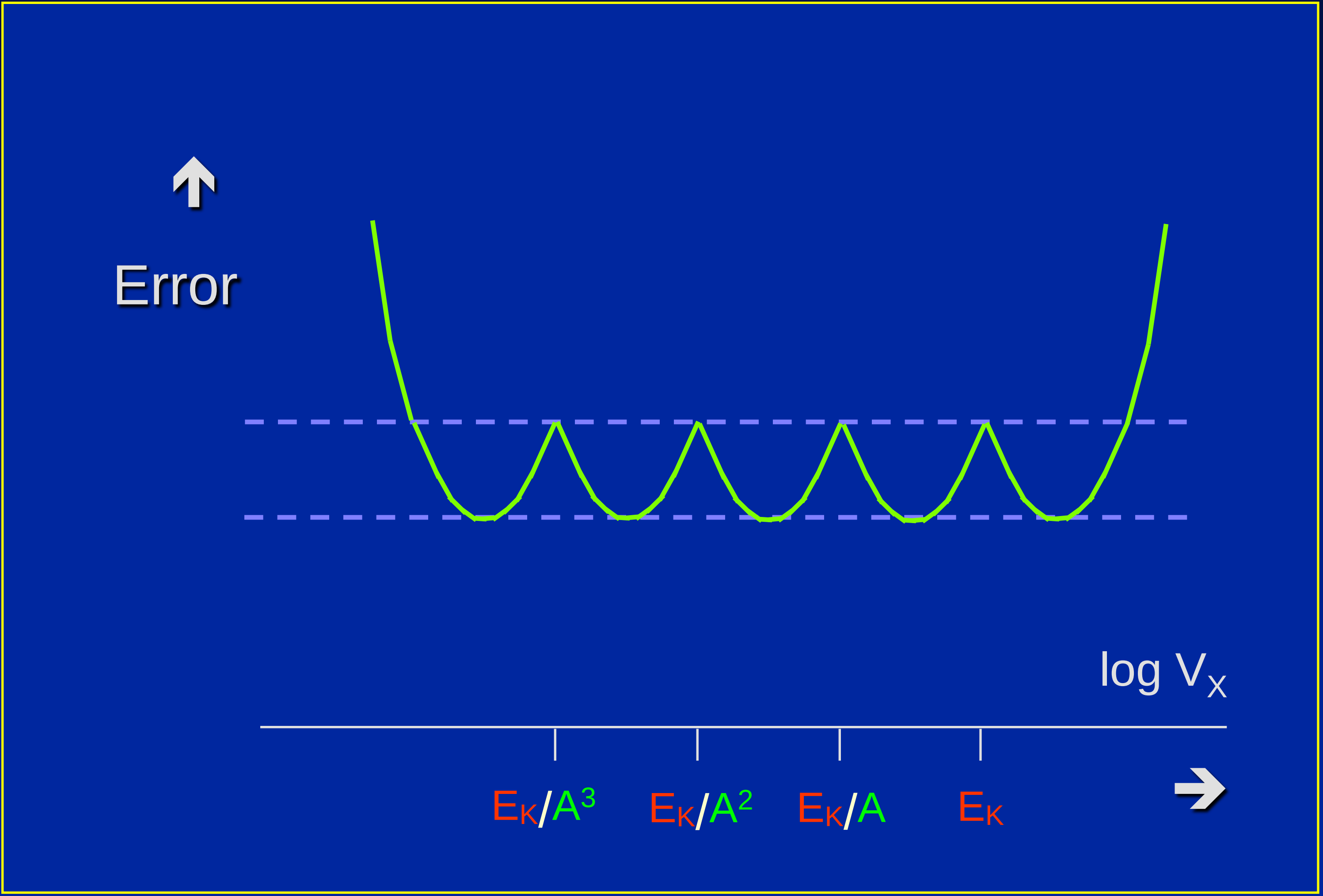
APPROXIMATION ERROR



LAW CONFORMANCE

dB Error = Actual Output (in dB)
minus Ideal Output (in dB)

Since there are 20 decibels per decade the ideal output in dB is

$$IDEAL_{dB} = 20 \lg t (V_X/V_Z)$$

LAW CONFORMANCE

The Actual Output in dB is

$$\text{ACTUAL}_{dB} = 20 \; \frac{\text{ACTUAL OUTPUT } V_W}{\text{VOLTS/DECADE } (V_Y)}$$

$$\text{ERROR}_{dB} = 20 \left\{ V_W/V_Y - \text{lgt}\,(V_X/V_Z) \right\}$$

LAW CONFORMANCE

For a log-amp based on the A/1 gain stages, the peak deviation from an ideal logarithmic response is

$$R_{dB} = 10\{(A + 1 - 2\sqrt{A})\,\mathrm{lgt}\,A\}/(A - 1)$$

Examples:

$A = 2$ (6dB),	$R = 0.52$dB;
$A = \sqrt{10}$ (10dB),	$R = 1.4$dB
$A = 4$ (12dB),	$R = 2.01$dB;
$A = 5$ (14dB),	$R = 2.67$dB

ANALOG DEVICES

DYNAMIC RANGE

THE FUNCTION FIT CONTINUES TO BE USEFULLY ACCURATE EVEN AT INPUTS BELOW THE FIRST TRANSITION AND ABOVE LAST ONE. IN PRACTICE, THE DYNAMIC RANGE IS SLIGHTLY MORE THAN A^N. THUS, USING A=4, N=10 WE COULD EXPECT A 120dB RANGE ......

DYNAMIC RANGE

... EXCEPT FOR THE LITTLE MATTER OF NOISE. FOR EXAMPLE, A FIRST-STAGE NOISE-SPECTRAL-DENSITY OF 1nV/√Hz IN A 500-MHZ BANDWIDTH IS EQUIVALENT TO AN INPUT-REFERRED NOISE OF 22.4μV RMS, OR -80dBm/50Ω

DYNAMIC RANGE

THE INPUT NEEDS TO BE AT LEAST 6dB ABOVE THE NOISE FLOOR TO MAKE AN ACCURATE MEASUREMENT, THUS AT -74dBm IN THIS EXAMPLE.

120dB ABOVE THIS NOISE FLOOR IS AT +46dBm (63V SINE AMPLITUDE!)

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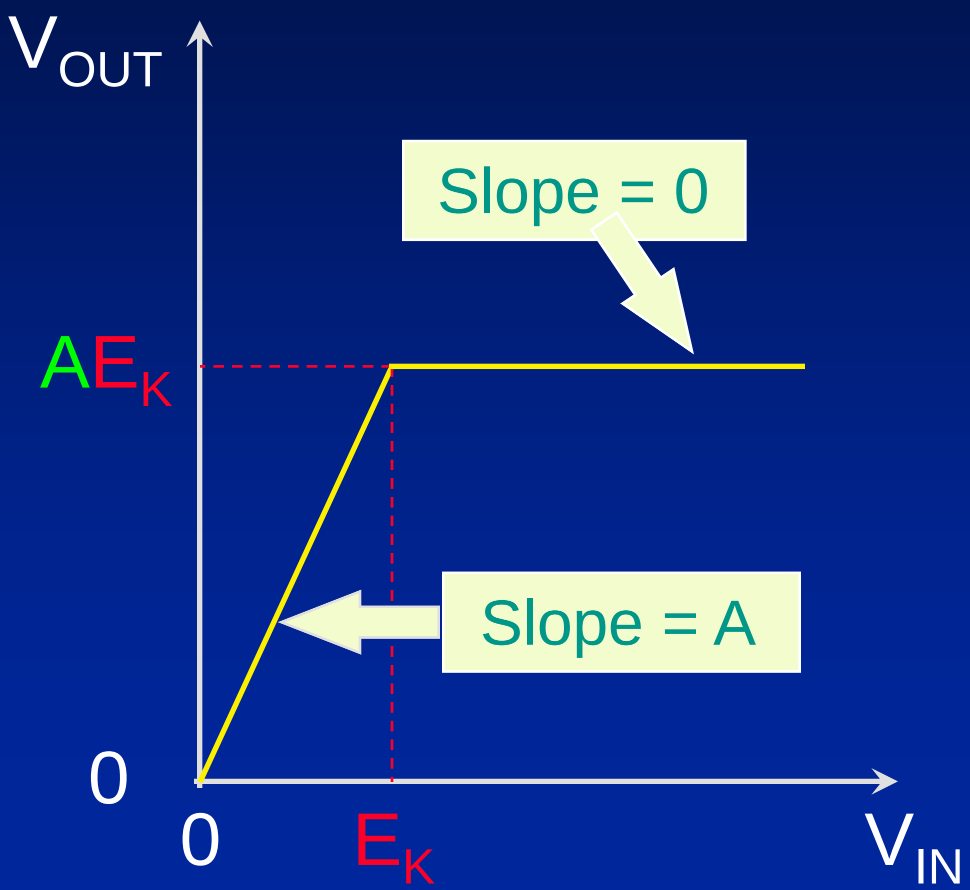
THE A/0 AMPLIFIER



THE A/0 AMPLIFIER

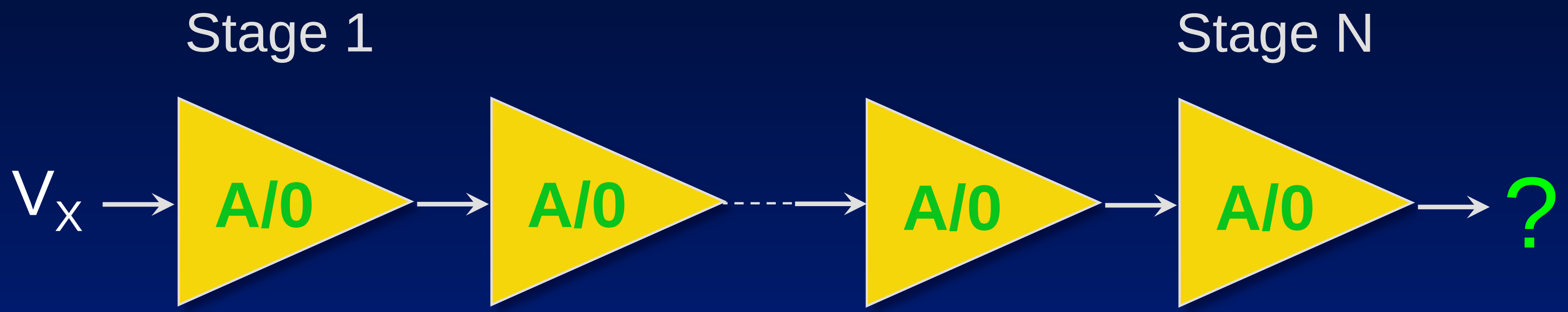
$V_{OUT} = AV_{IN}$

for $V_{IN} < E_K$

$V_{OUT} = AE_K$

for $V_{IN} > E_K$

N-CASCADE OF A/0 CELLS



WE CAN NO LONGER TAKE THE OUTPUT FROM THE LAST STAGE: IT LIMITS TO AE_K AS SOON AS ITS INPUT IS E_K AND THEREAFTER *IT DOES NOT INCREASE*

N-CASCADE OF A/0 CELLS

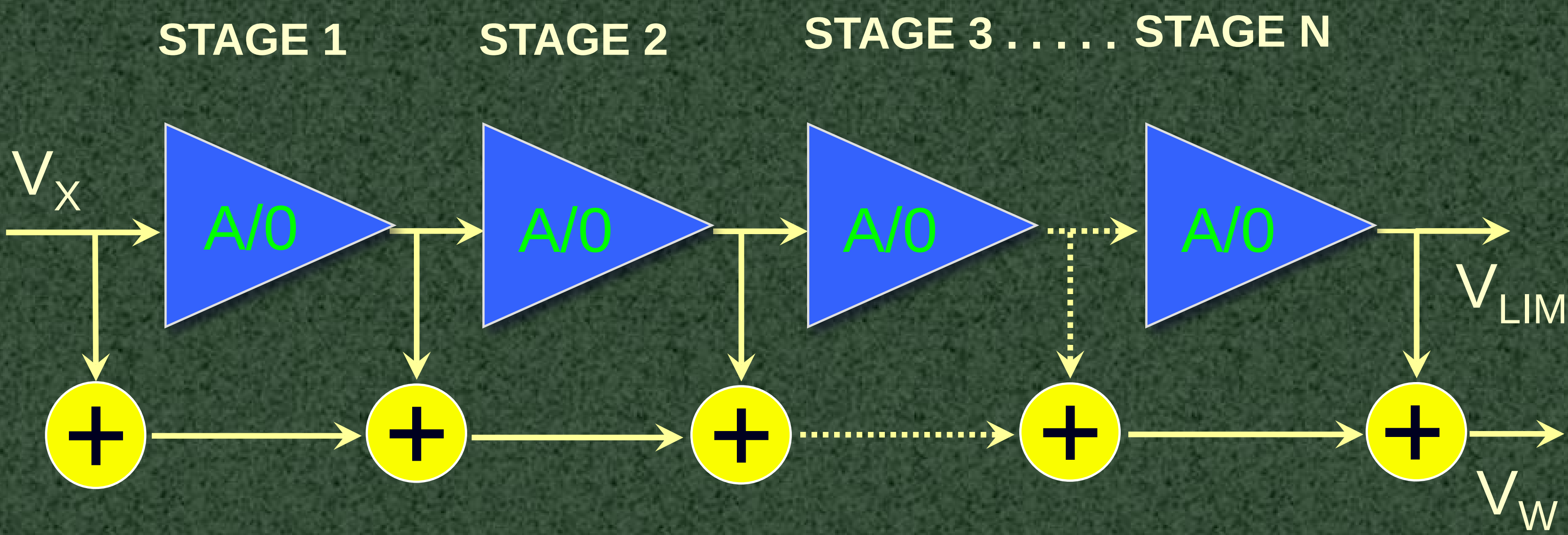
WHILE WE MUST STILL FIND

$$V_Y = y E_K$$

$$V_Z = z E_K$$

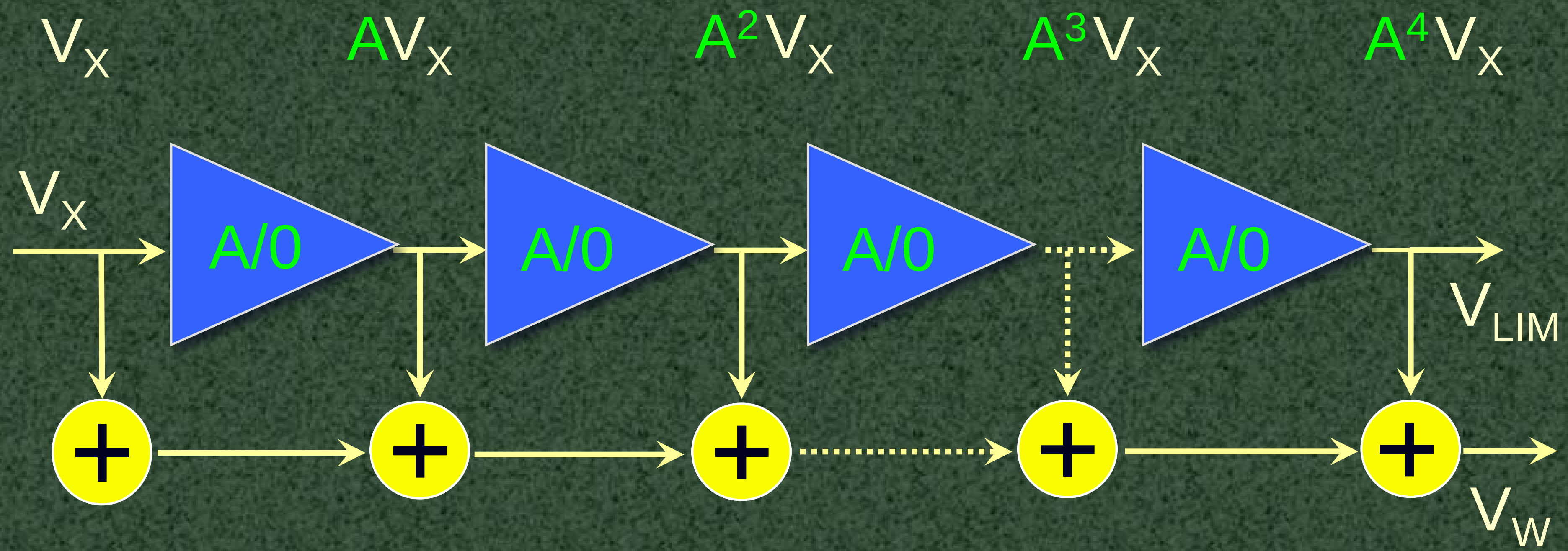
we should expect y and z to now be different functions A and N than for the cascade of A/1 stages

THE SOLUTION: ADD ALL THE OUTPUTS TOGETHER



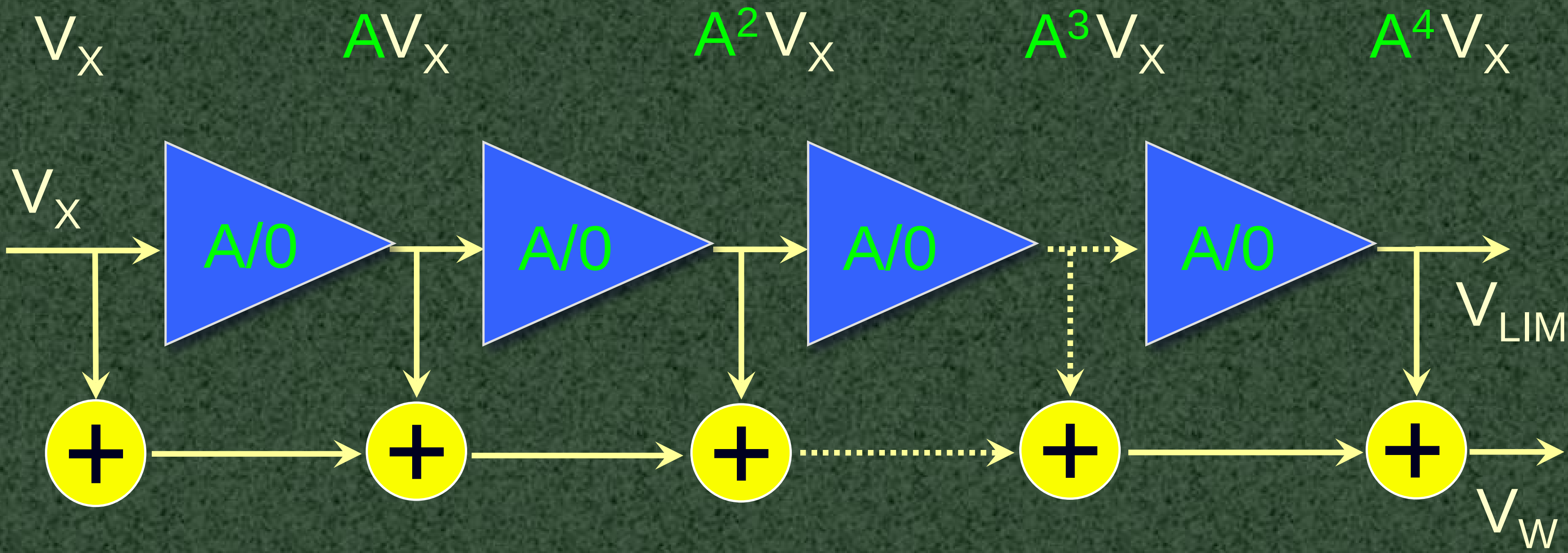
THIS IS THE BASIC BACKBONE OF PRACTICALLY ALL MONOLITHIC PROGRESSIVE-COMPRESSION LOG-AMPS; NOTE THAT LAST STAGE OUTPUT IS NOW CALLED V_{LIM}

FOUR-STAGE EXAMPLE



- INPUT VOLTAGE VERY SMALL
- ALL STAGES REMAIN LINEAR
- INTERNAL VOLTAGE E_K HIDDEN

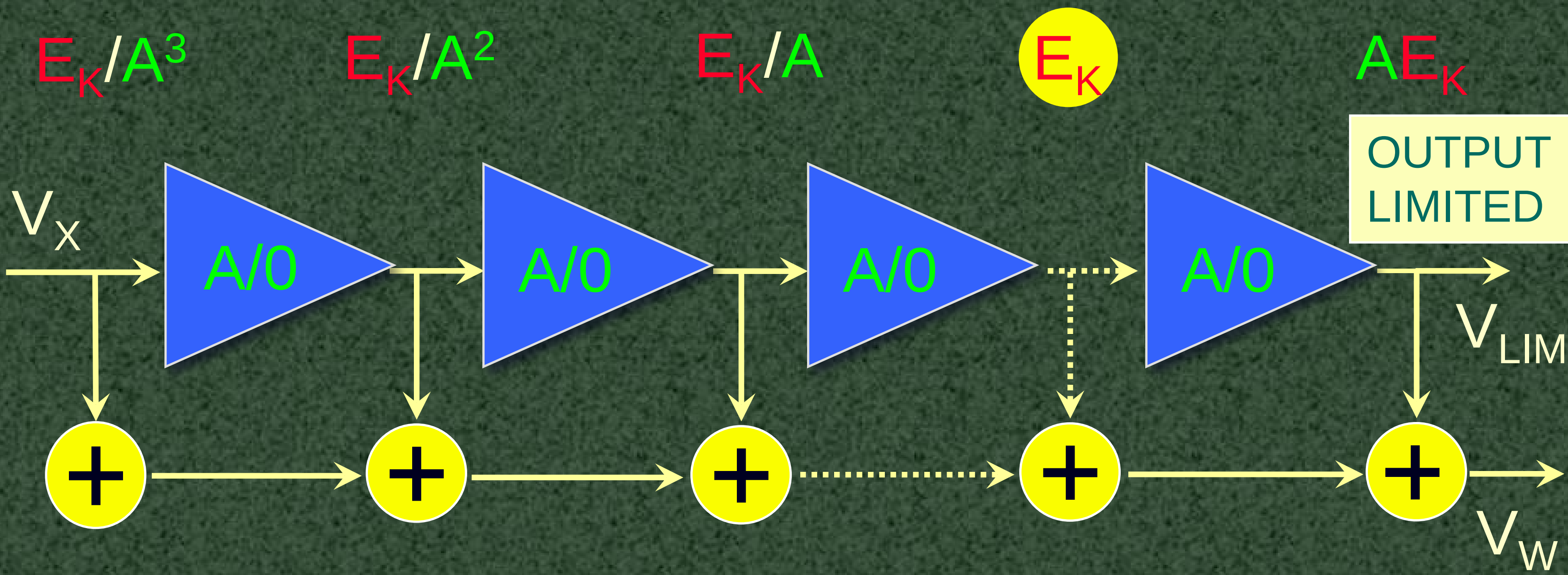
FOUR-STAGE EXAMPLE



FOR SMALL INPUTS

$$V_W = (1 + A + A^2 + A^3 + A^4)V_X$$

FOUR-STAGE EXAMPLE



FOR AN INPUT OF EXACTLY E_K/A^3

$$V_W = (A+1+A^{-1}+A^{-2}+A^{-3})E_K$$

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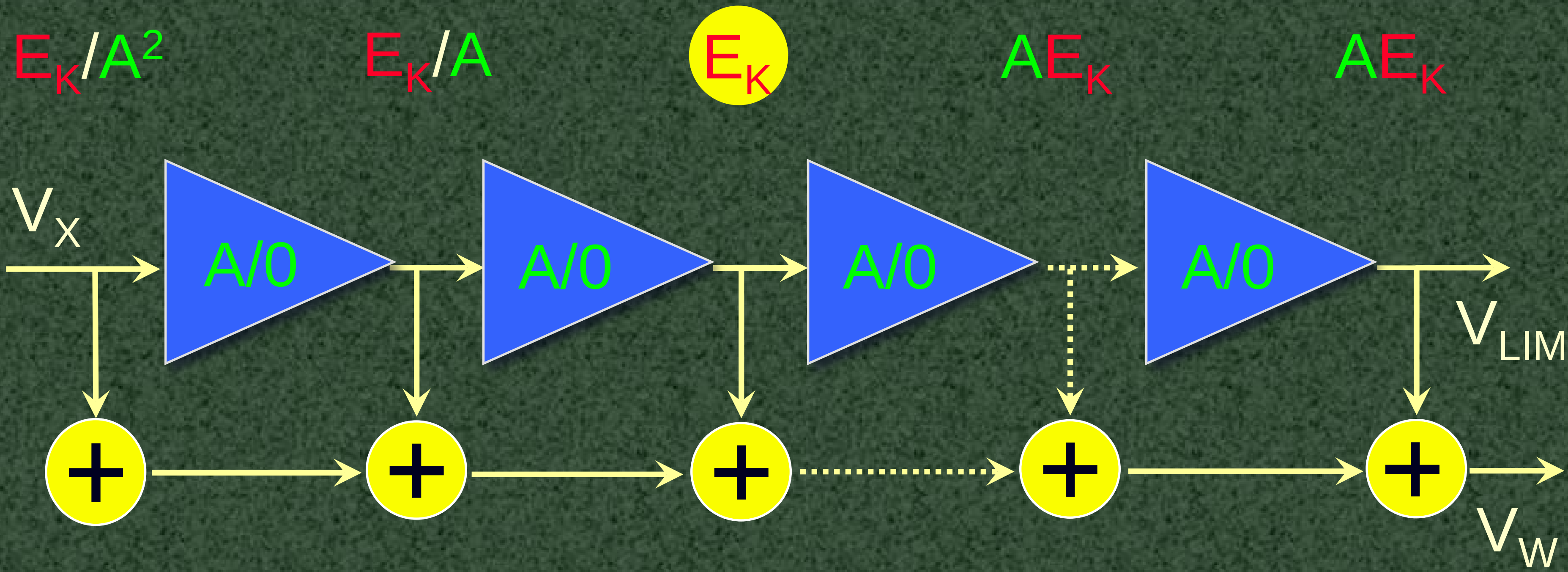
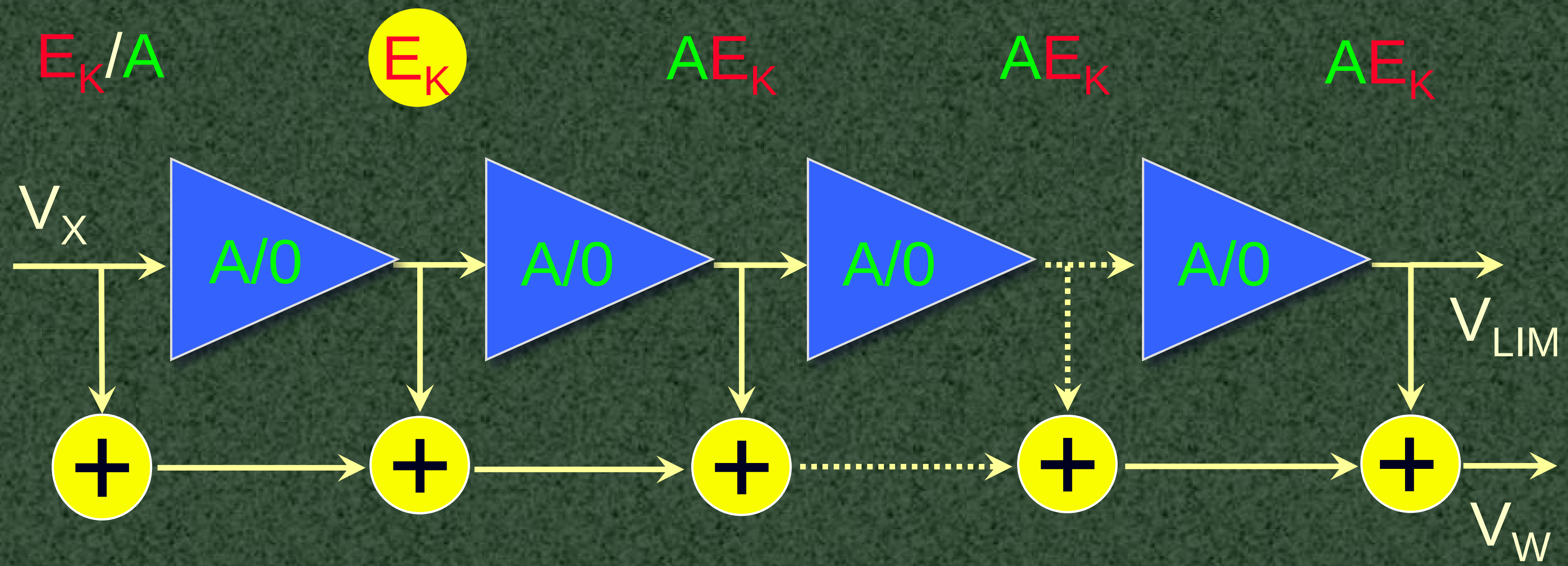
FOUR-STAGE EXAMPLE



FOR AN INPUT OF EXACTLY E_K/A^2

$$V_W = (2A + 1 + A^{-1} + A^{-2})E_K$$

FOUR-STAGE EXAMPLE



FOR AN INPUT OF EXACTLY E_K/A

$$V_W = (3A+1+A^{-1})E_K$$

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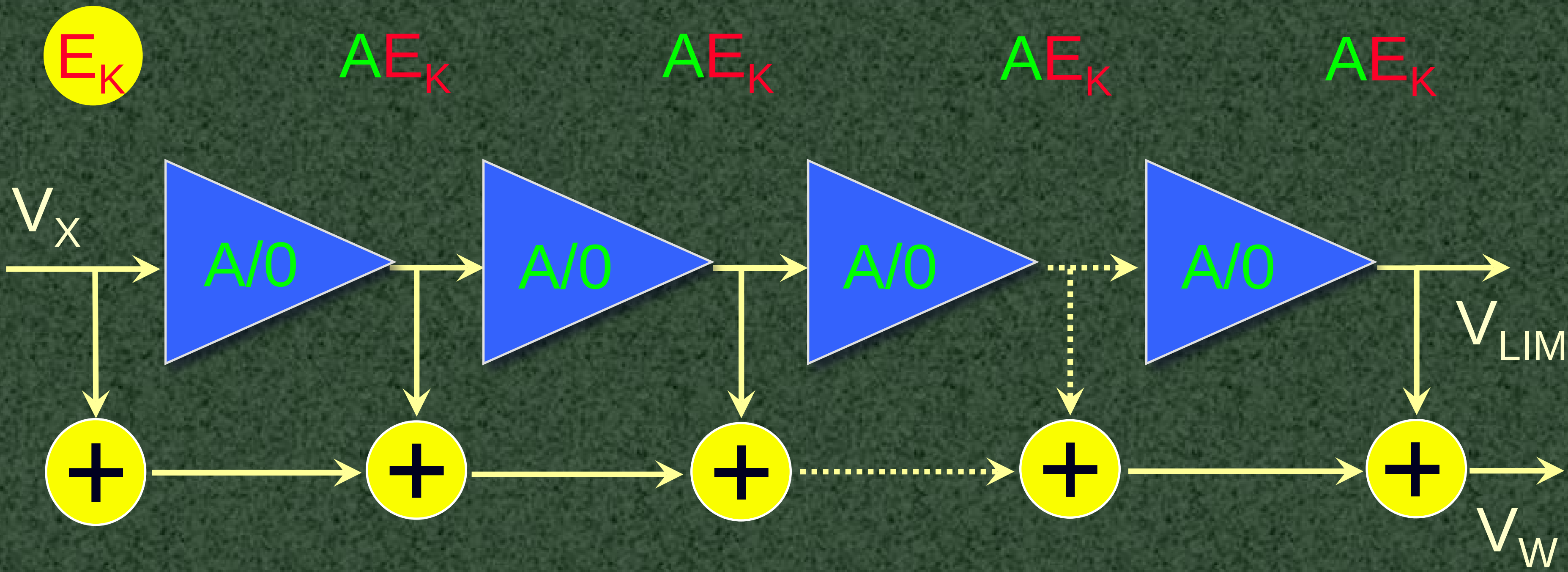
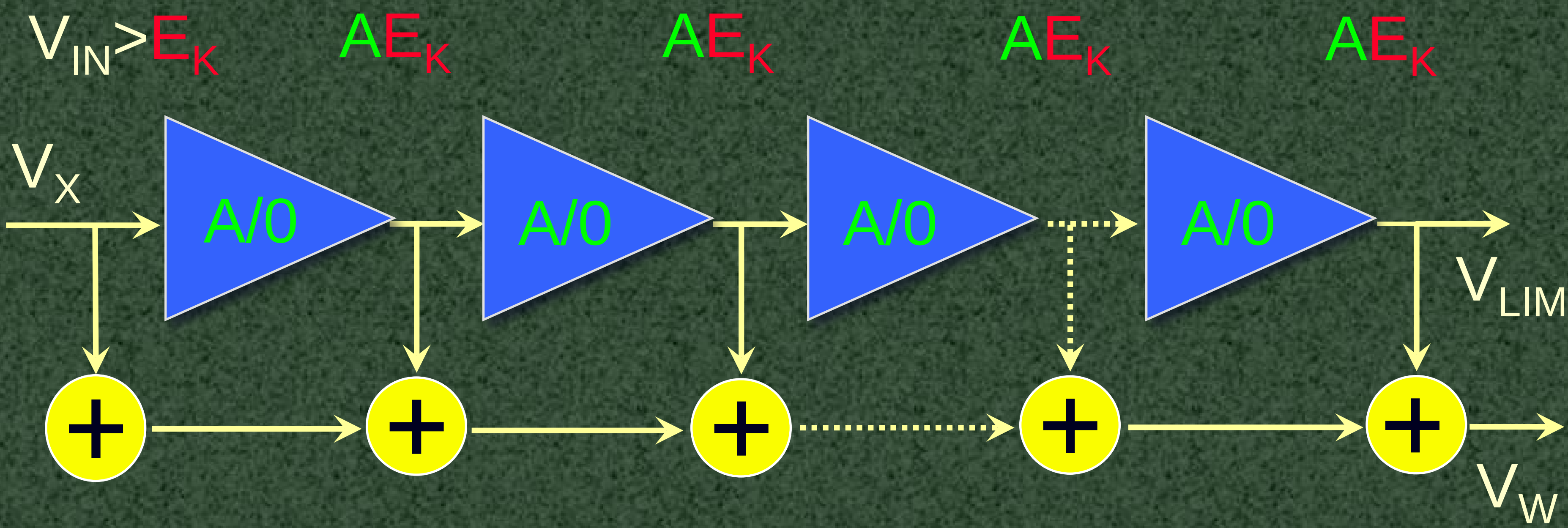
FOUR-STAGE EXAMPLE



FOR AN INPUT OF EXACTLY E_K

$$V_W = (4A+1)\,E_K$$

FOUR-STAGE EXAMPLE



FOR ANY INPUT ABOVE E_K

$$V_W = 4AE_K + V_{IN}$$

SLOPE CALCULATION

THE OUTPUT CHANGES BY essentially* AE_K as the INPUT CHANGES BY THE *RATIO* A

$$V_Y = \frac{AE_K}{\mathrm{lgt}\,(A)} \quad \text{(volts/decade)}$$

* it's actually $A\,(1 - A^{-N})E_K$

SLOPE CALCULATION

IT IS INTERESTING TO NOTE

THAT THE FUNCTION A/lgt(A)

VARIES BY ONLY 6% OVER

$2 < A < 4$ ($6.64 \rightarrow 6.26 \rightarrow 6.64$)

WITH ITS MINIMUM AT $A = e$

SLOPE CALCULATION

SO, BY CHOOSING TO USE A GAIN OF A = *e* (2.72, or 8.7dB) THE LOG SLOPE WILL EXHIBIT *ZERO SENSITIVITY* TO SMALL VARIATIONS IN ACTUAL GAIN: IT WOULD BE SIMPLY 6.26E_K

INTERCEPT FOR A/0 SYSTEM

USING THE SAME APPROACH AS FOR THE A/1 CASE, IT IS FOUND THAT THE INTERCEPT IS POSITIONED AT

$$V_Z = \frac{E_K}{A^{N + 1/(A-1)}}$$

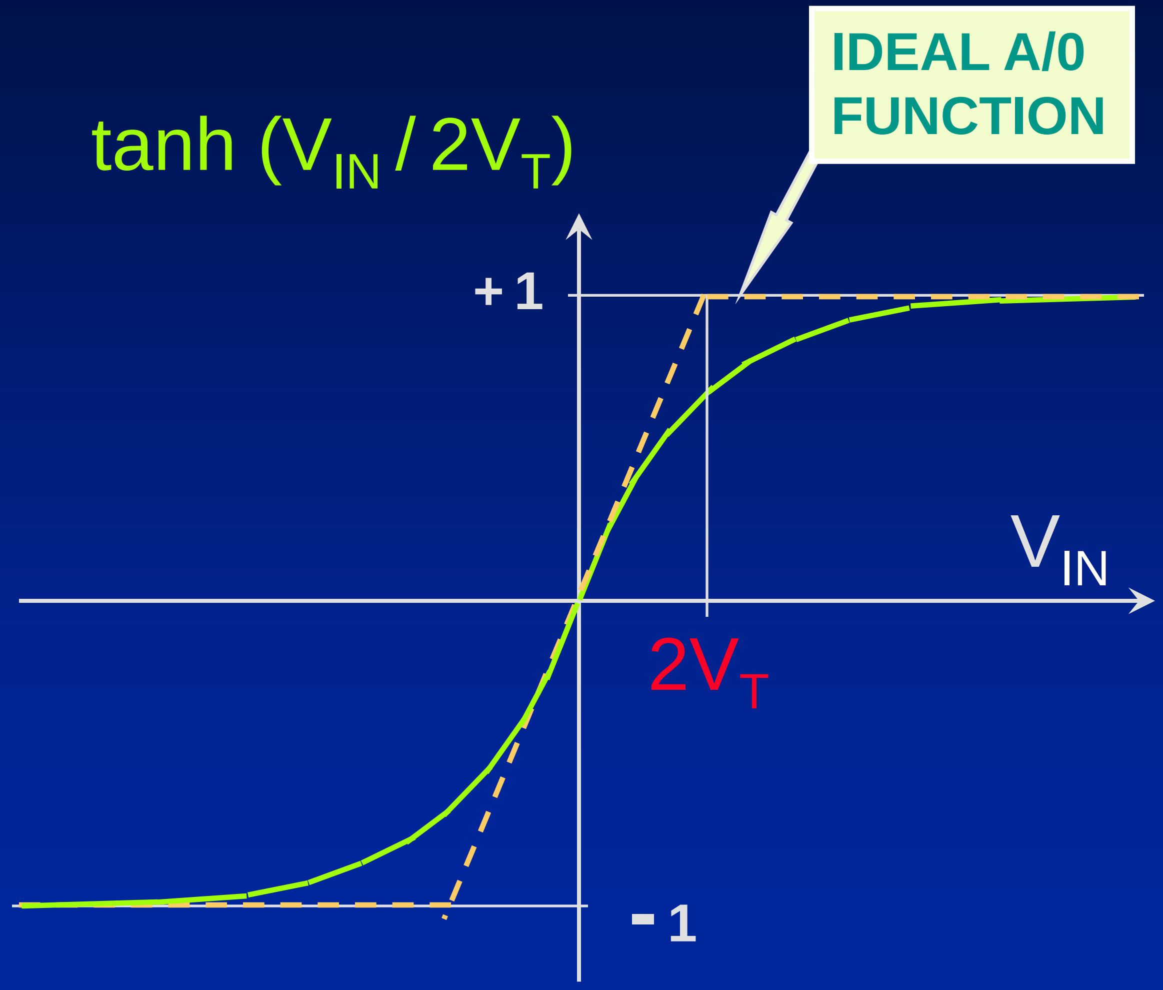
PRACTICAL REALIZATION

Progressive-compression log amps have been implemented in many different technologies. In the past this included discrete and monolithic BJT and GaAs embodiments.

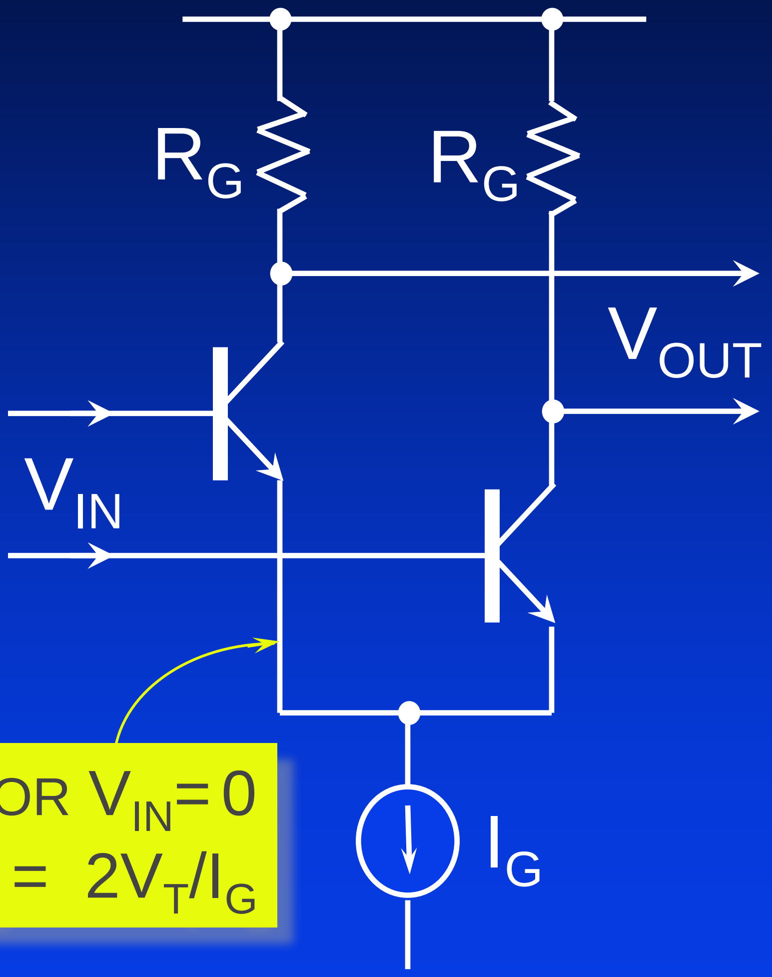
Today, one can use deep sub-micron CMOS to build log amps for operation up to several GHz.

In this presentation, we'll illustrate some design techniques using monolithic bipolar processes.

tanh as LIMITER



An A/0 AMPLIFIER USING tanh



FOR $V_{IN}=0$
$r_e = 2V_T/I_G$

LARGE-SIGNAL FUNCTION:

$V_{OUT} = I_G R_G \tanh(V_{IN}/2V_T)$

SMALL-SIGNAL GAIN (V_{IN}=0)

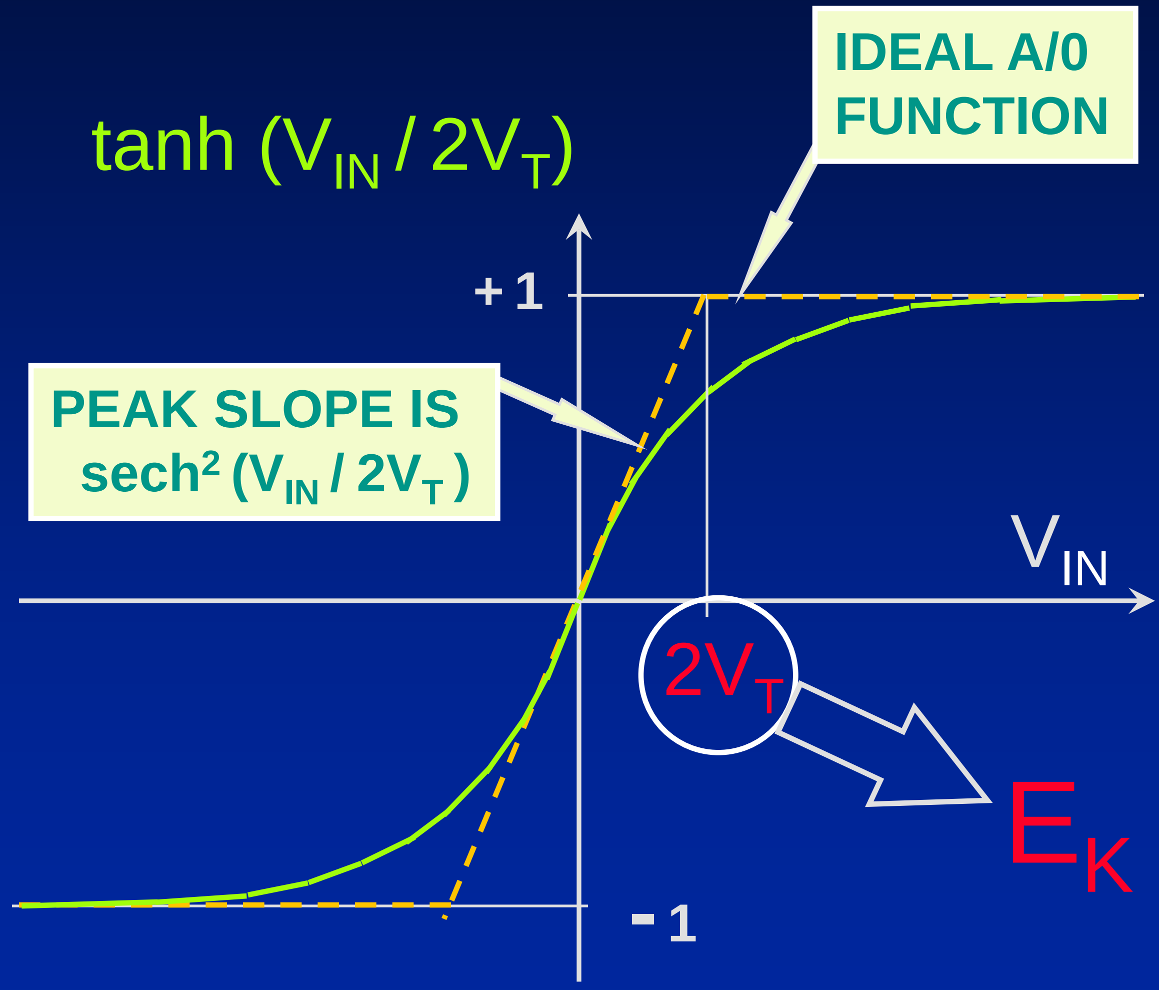
$A_0 = R_L/r_e = I_G R_G/2V_T$

INCREMENTAL GAIN

$\partial V_{OUT}/\partial V_{IN} = A_0 \operatorname{sech}^2(V_{IN}/2V_T)$

where $V_T = kT/q$

tanh as LIMITER



ANALOG DEVICES

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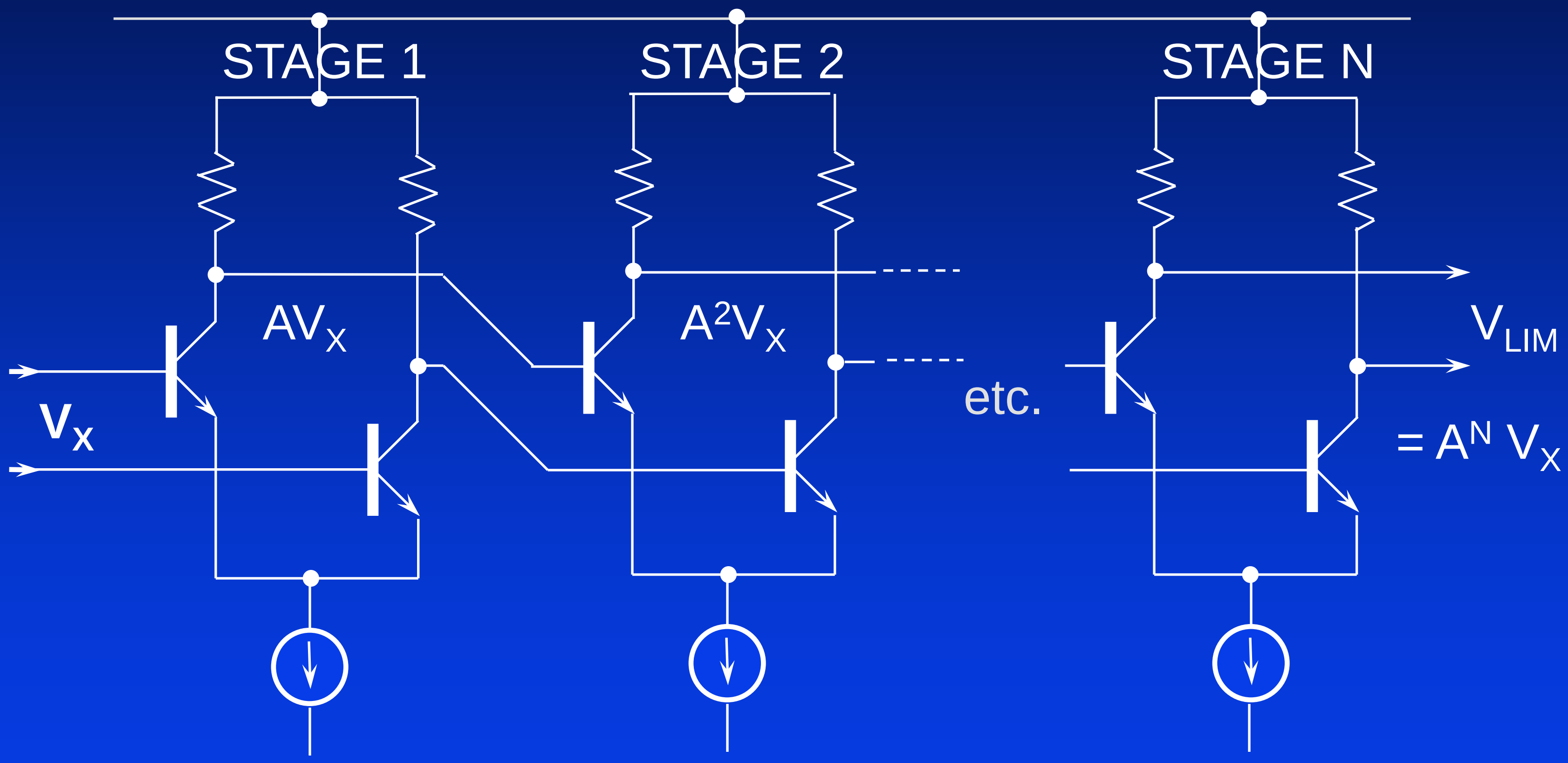
CASCADE OF tanh AMPLIFIER CELLS



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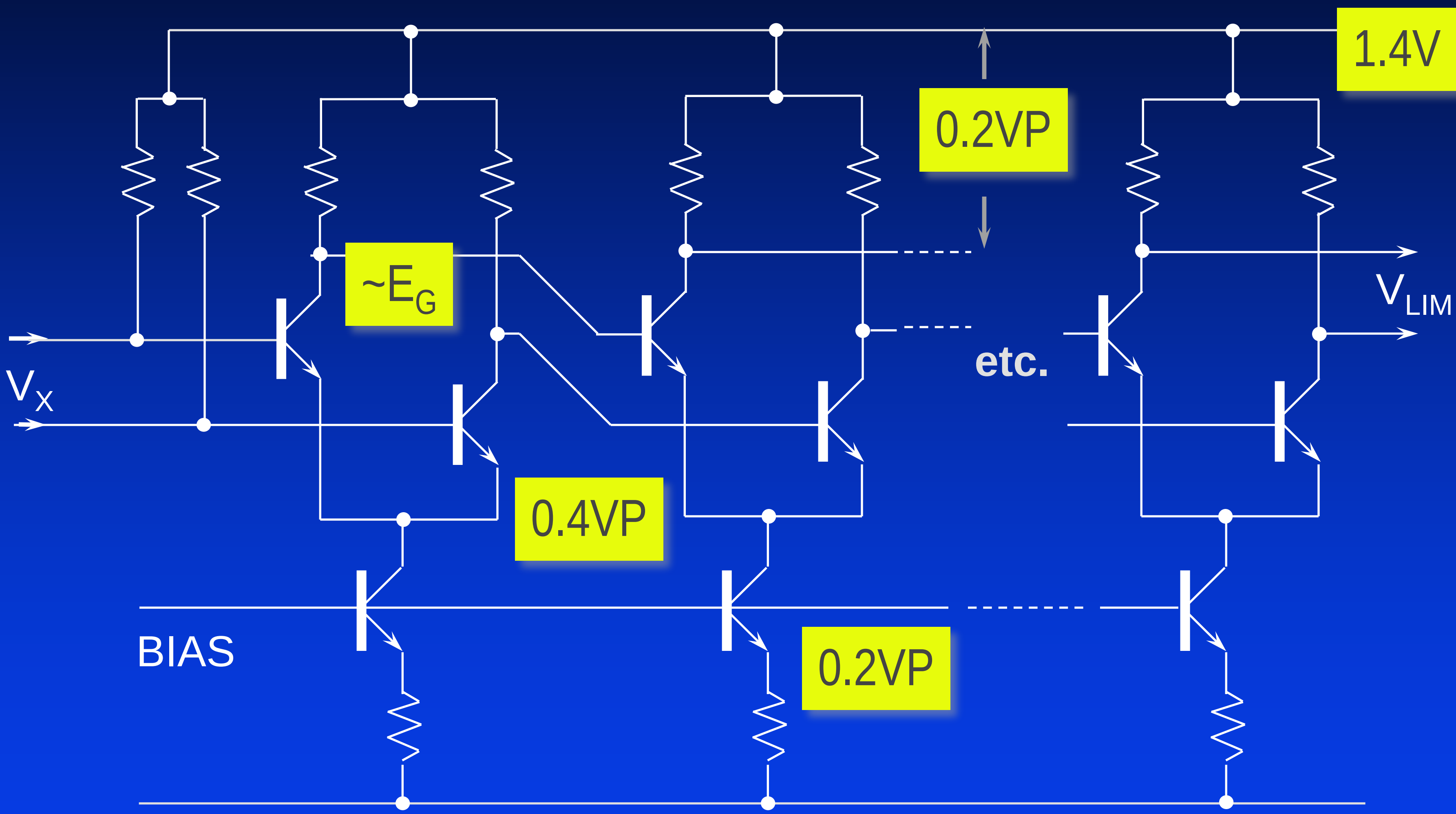
LOW SUPPLY VOLTAGE



BUT AS ALWAYS IN PRACTICE, THE DEVIL'S IN THE DETAILS….

…. SCORES OF THEM ….

SUCH AS THE EFFECTS OF

FINITE DC AND AC BETA;

CHOICE OF "T-SHAPES";

OHMIC RESISTANCES;

FINITE INERTIA; etc.

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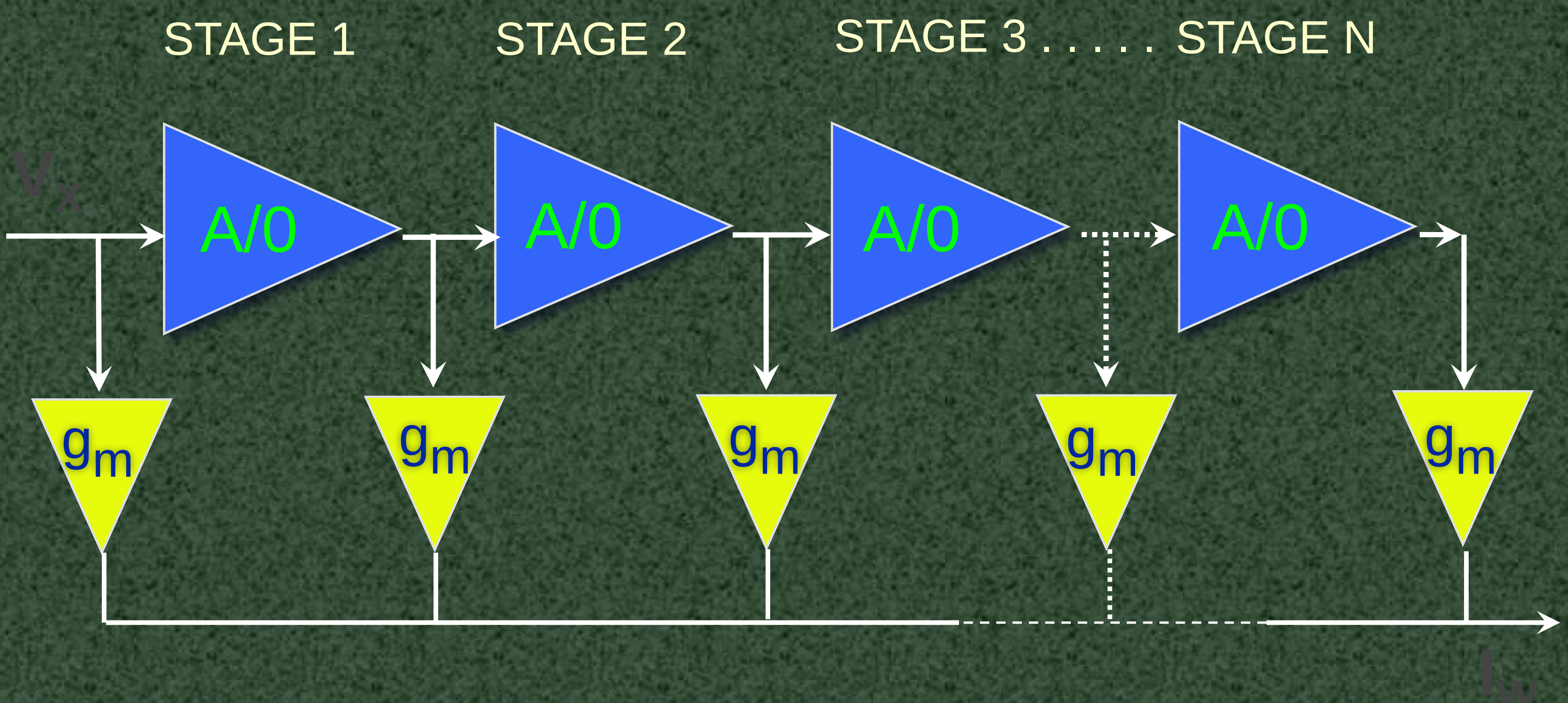
AUXILIARY g_m CELLS SUM IN CURRENT-MODE



USEFUL SECONDARY FUNCTION: gm CELLS ISOLATE GAIN STAGES

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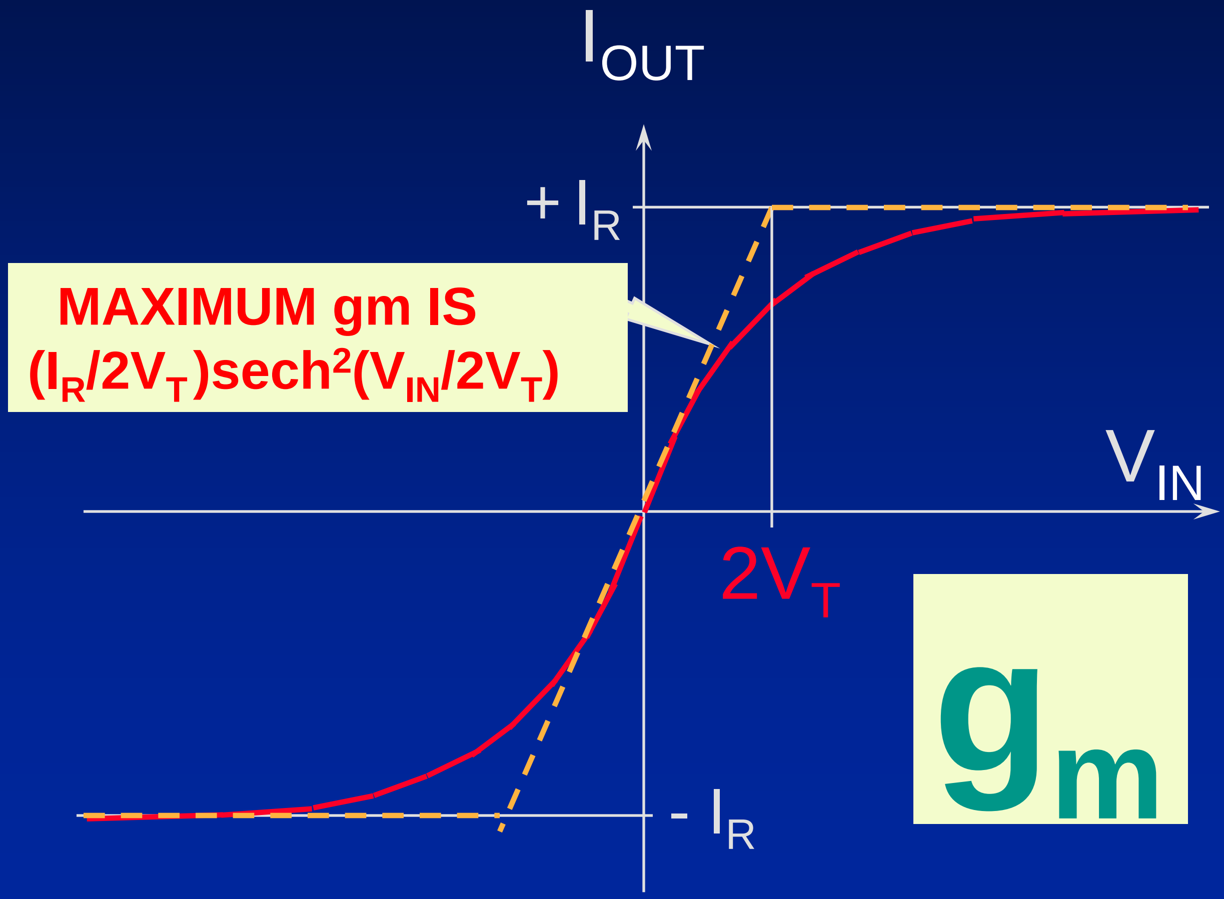
AUXILIARY gm CELLS: STILL USE THE A/0 IDEA



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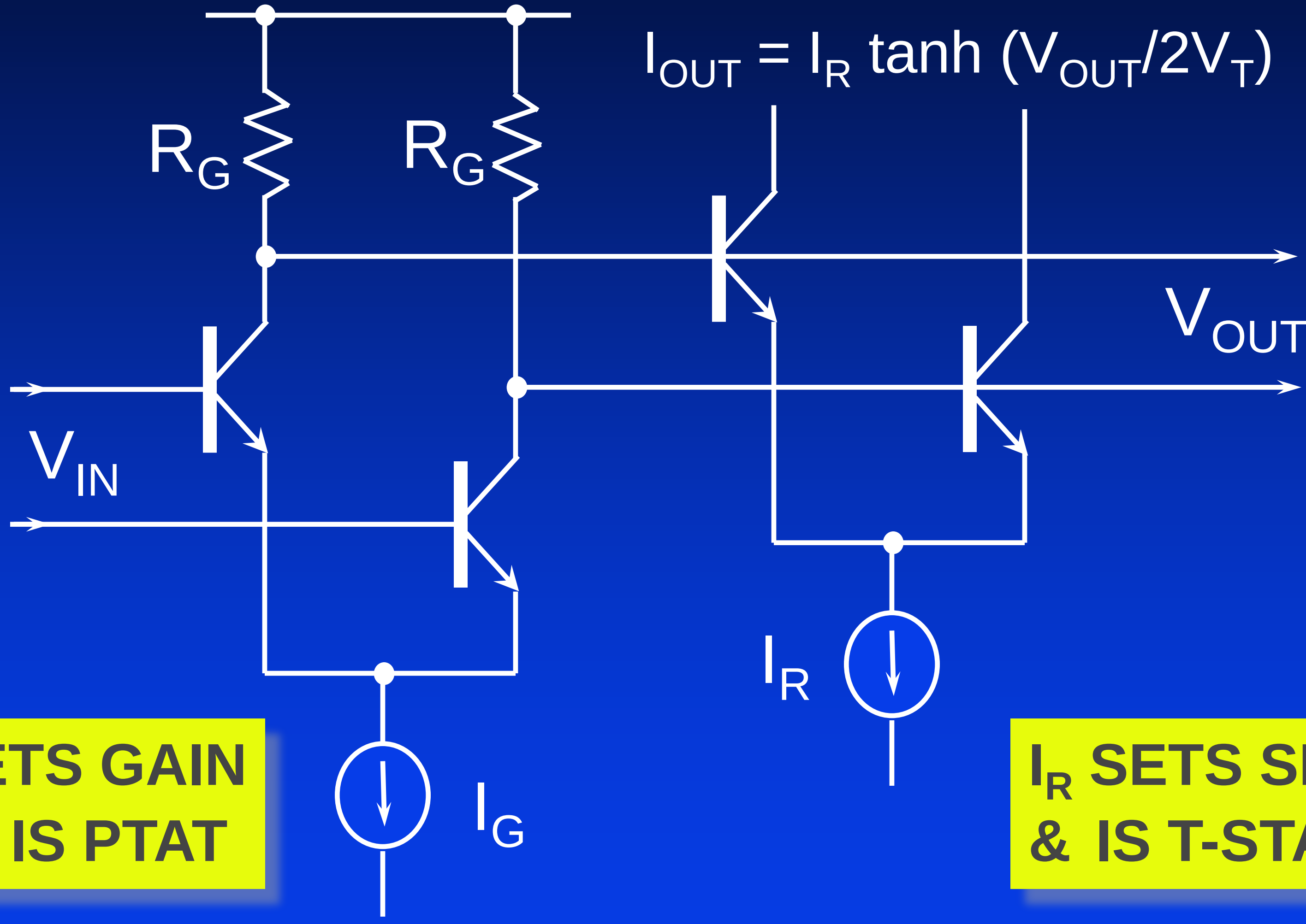
AUXILIARY g_m CELLS

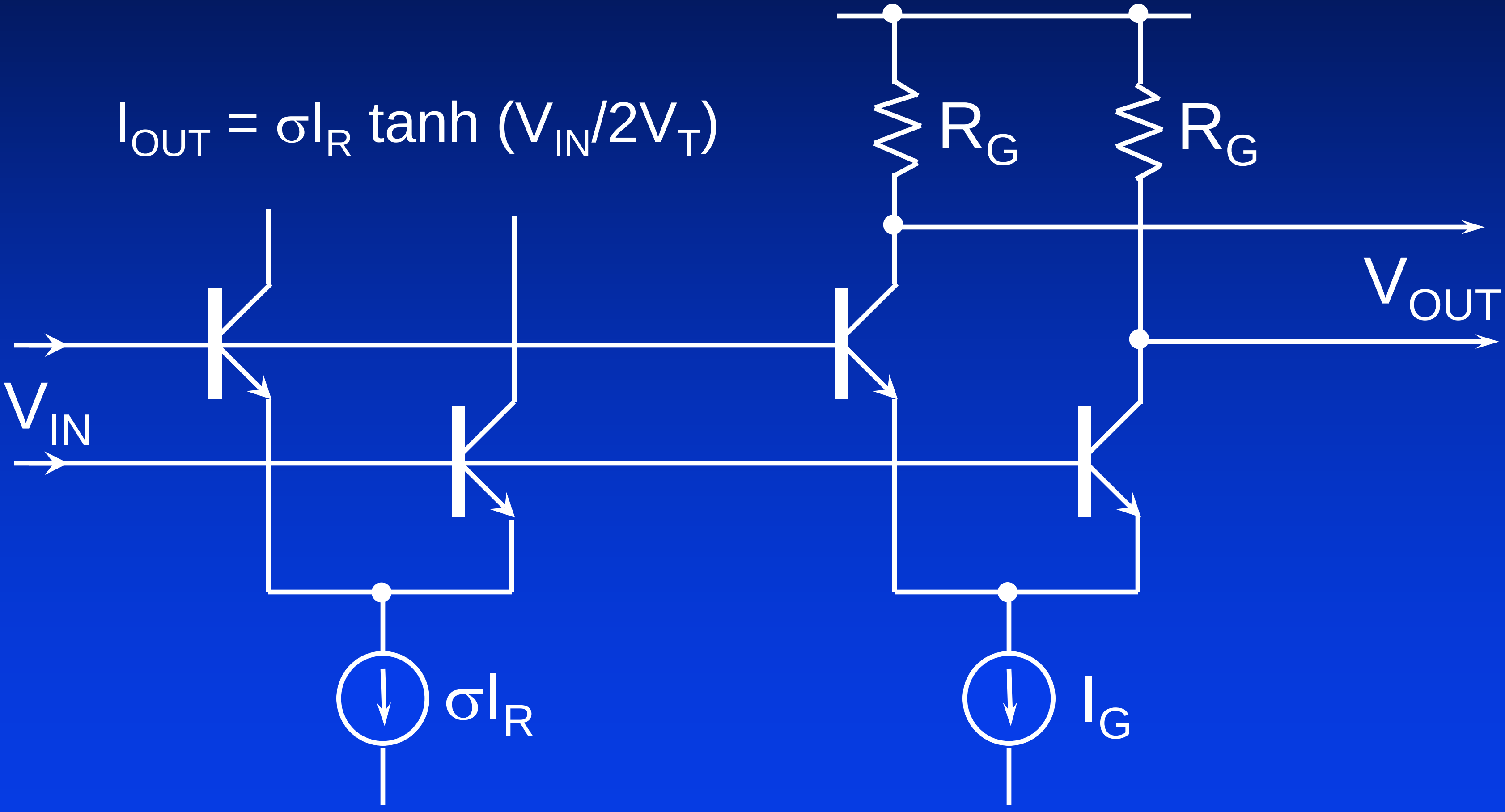


I_G SETS GAIN AND IS PTAT

I_R SETS SLOPE & IS T-STABLE

DIFFERENT FIRST STAGE

$I_{OUT} = \sigma I_R \tanh(V_{IN}/2V_T)$



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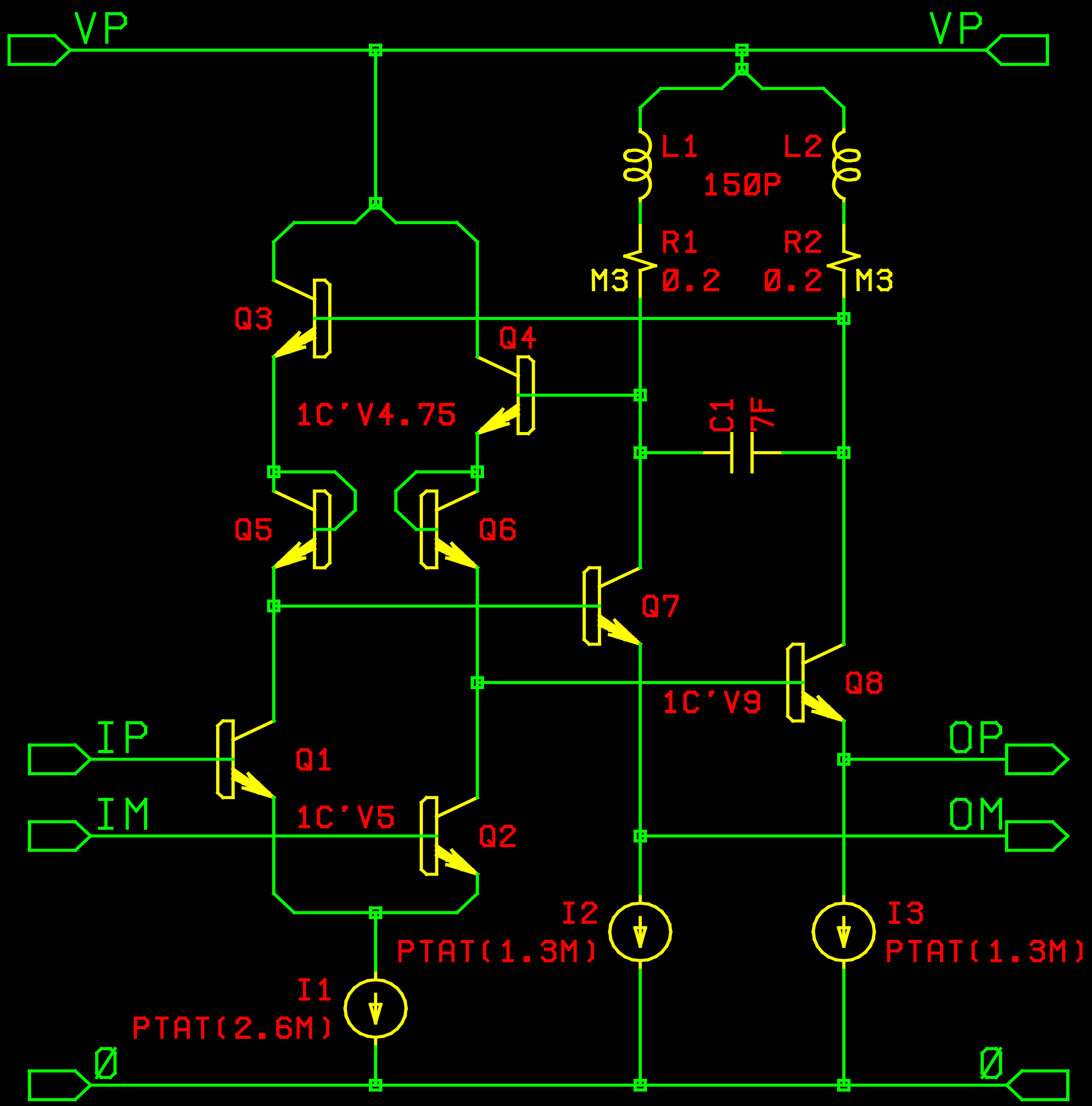
A TRANSLINEAR AMPLIFIER



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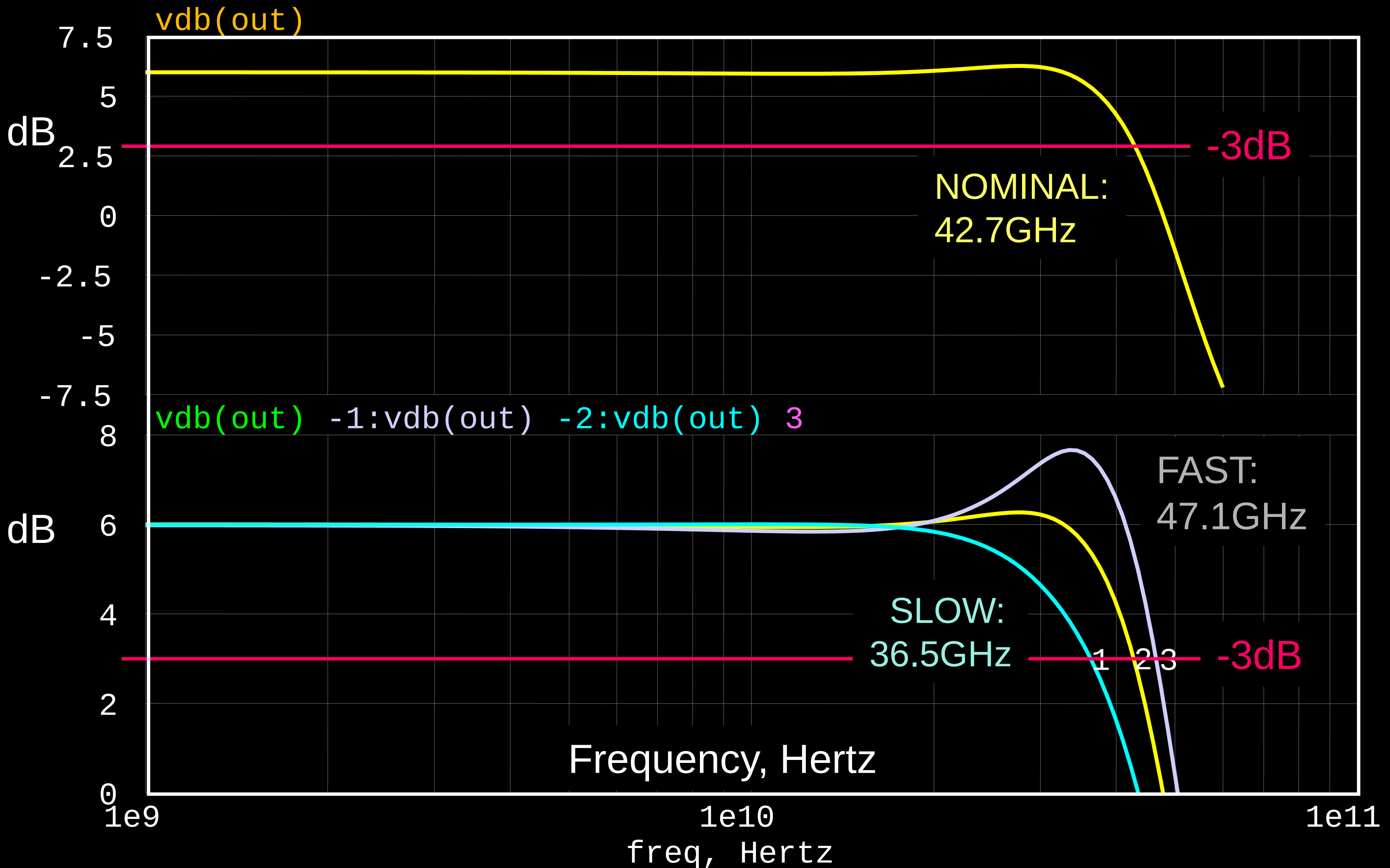
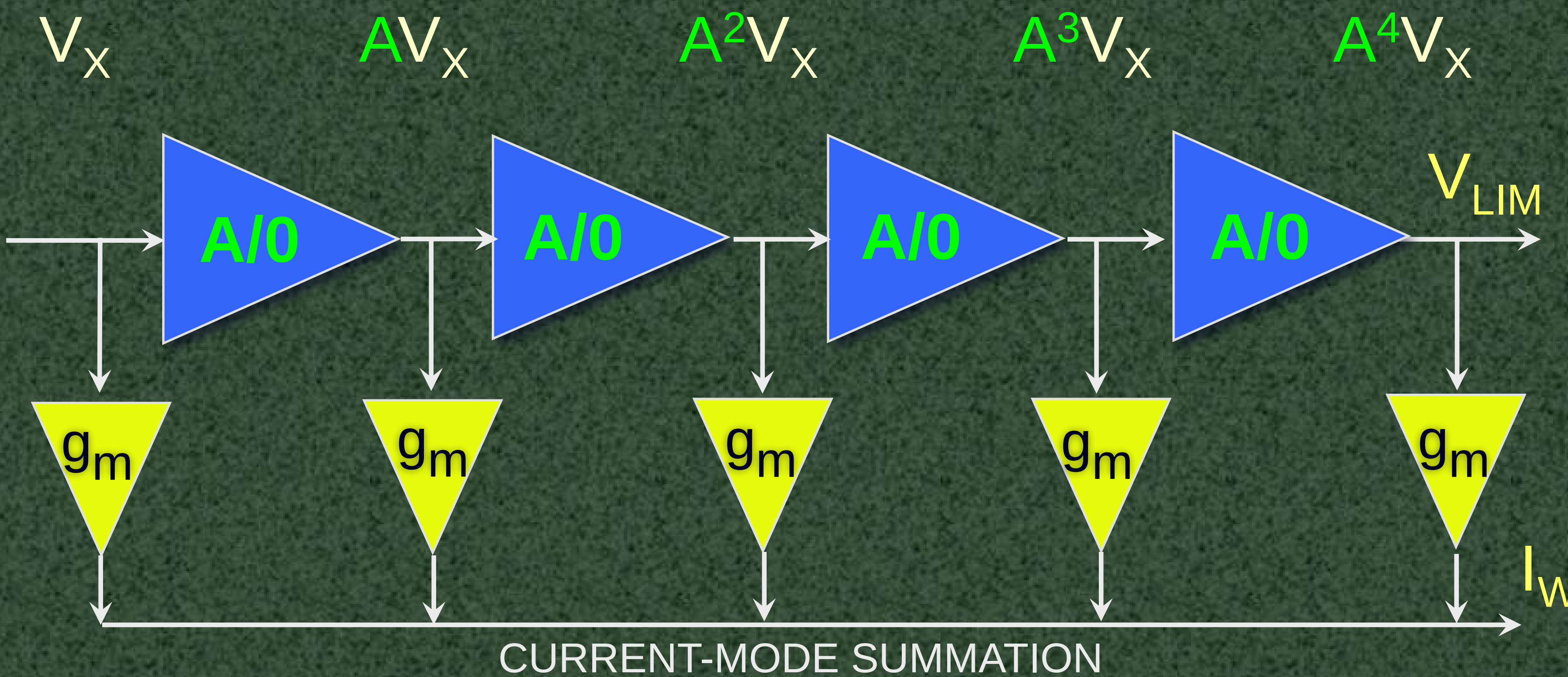
TEN CASCADED STAGES, 150pH INDUCTORS

CASCODES: VCS=2.28V, VCM=3.3V, VPS = 4.45V; EF's are 5um; Uppers = 4.75um
-cascade of ten stages with inductors included. f_3dB for the single stage is 42.7GHz, nominal, 47.1GHz FAST and 36.5GHz SLOW, sigma=3



A FOUR-STAGE EXAMPLE



FOR SMALL INPUTS, SYSTEM IS A LINEAR AMPLIFIER

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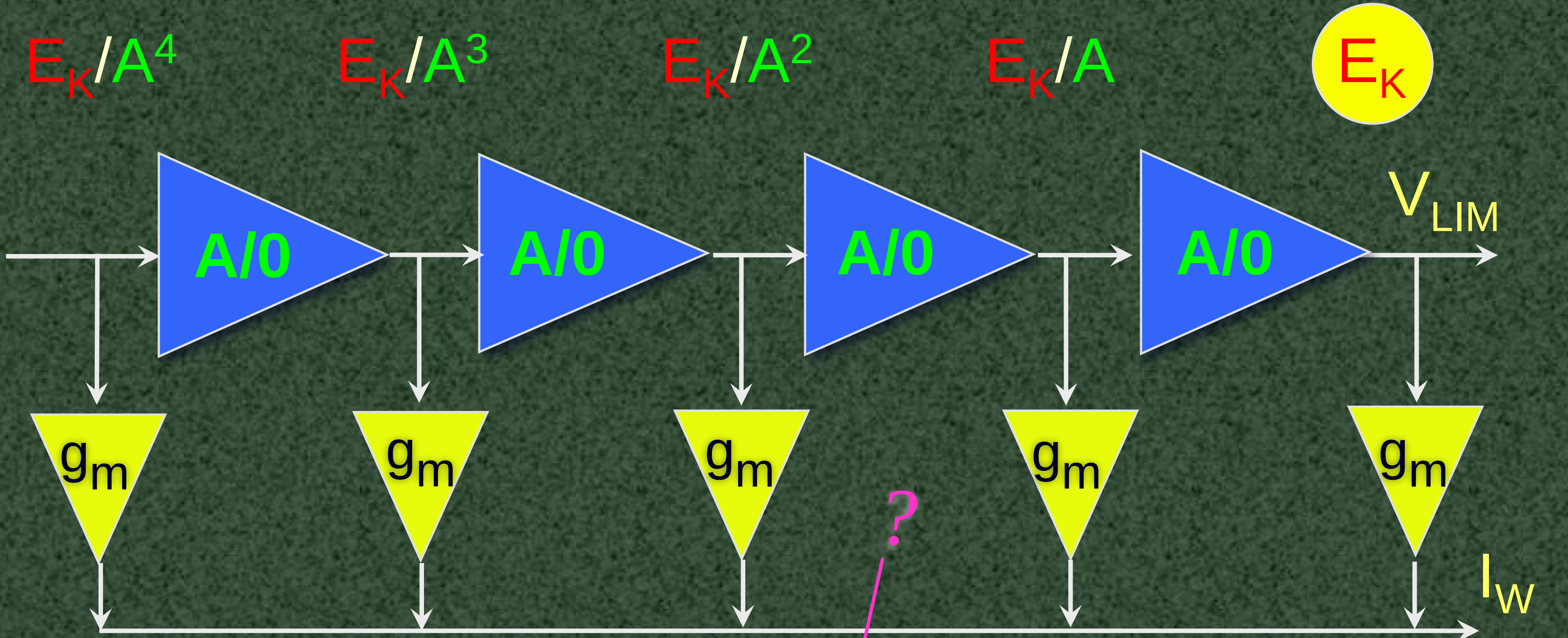
A FOUR-STAGE EXAMPLE



THE FIRST TRANSITION NOW OCCURS WHEN THE INPUT TO THE FINAL gm CELL REACHES E_K

AT THIS POINT THE INPUT IS E_K/A^4 AND THE OUTPUT CURRENT $I_W = \lambda I_R(1 + A^{-1} + A^{-2} + A^{-3} + A^{-4})$

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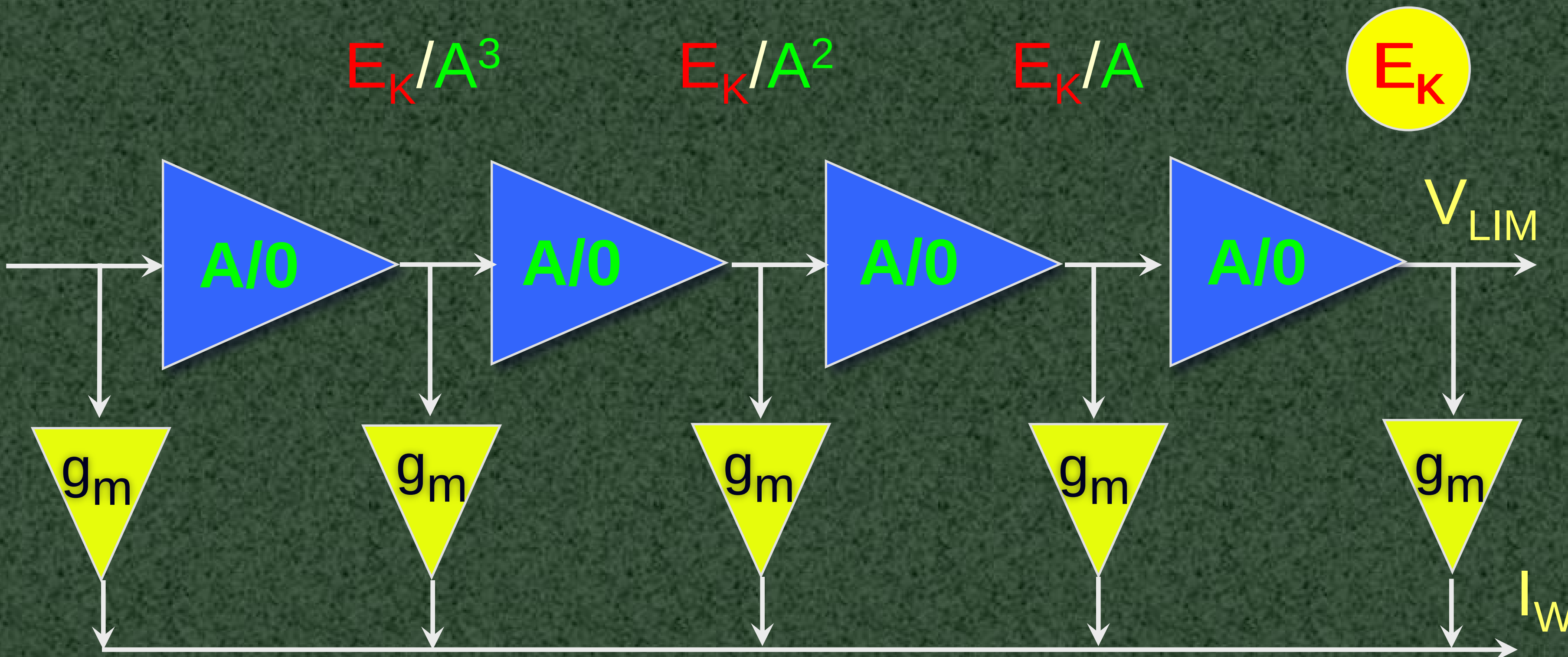
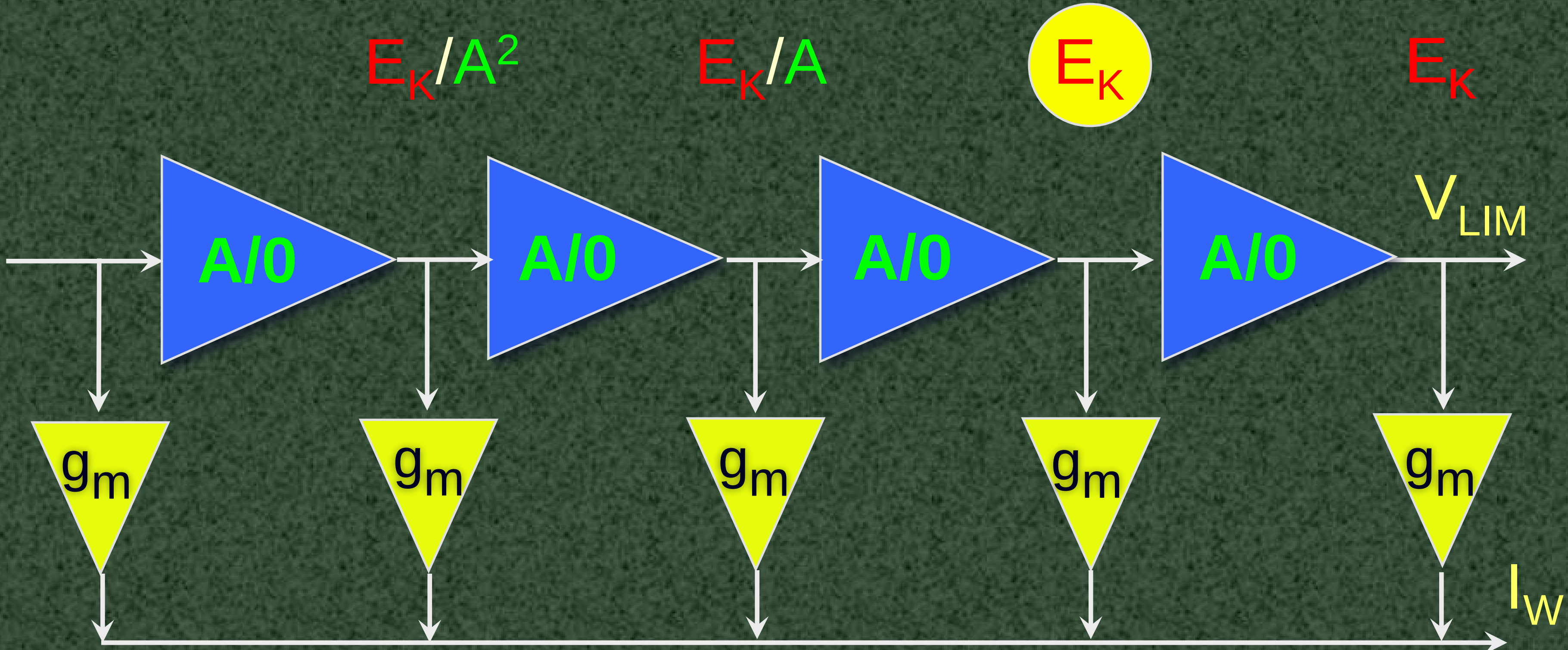
A FOUR-STAGE EXAMPLE



WHEN THE INPUT IS A TIMES HIGHER, AT E_K/A^3, THE OUTPUT IS $I_W = I_R(2 + A^{-1} + A^{-2} + A^{-3})$

A FOUR-STAGE EXAMPLE



WHEN THE INPUT IS A TIMES HIGHER, AT E_K/A^2,
THE OUTPUT IS $I_W = I_R(3 + A^{-1} + A^{-2} + A^{-3})$ etc.
etc.

DETERMINATION OF SLOPE

THE OUTPUT CHANGES

from $I_W = I_R (1 + A^{-1} + A^{-2} + \ldots\ldots. A^{-N})$

for an input of $V_X = E_K/A^{-N}$,

to $I_W = I_R (2 + A^{-1} + A^{-2} + \ldots\ldots. A^{-(N-1)})$

for an input of $V_X = E_K/A^{-(N-1)}$.

The output changes by $I_R(1 - A^{-N}) \cong I_R$

as the INPUT changes by the RATIO A

DETERMINATION OF SLOPE

WE NEED NOT PROCEED FURTHER: THE LOG FUNCTION DEVELOPS IN A SIMILAR FASHION TO THAT DETERMINED EARLIER FOR THE MORE IDEALIZED CASE.

THE LOG SLOPE – NOW IN CURRENT FORM – IS SIMPLY

$$I_W = \frac{I_R}{\mathrm{lgt}\,(A)}$$

WHICH IS FULLY DECOUPLED FROM E_K

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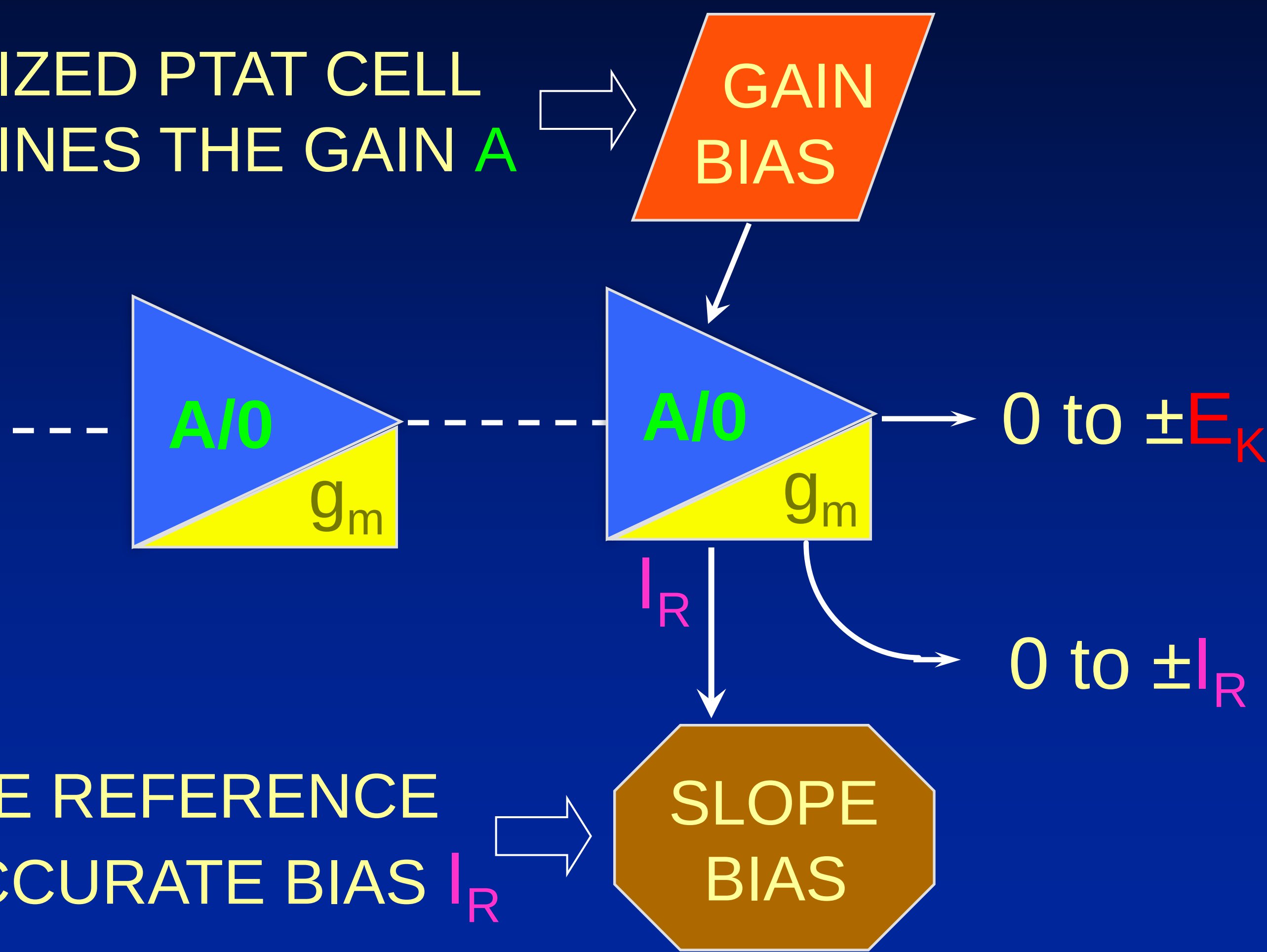
ELIMINATION OF PTAT E_K

SPECIALIZED PTAT CELL DETERMINES THE GAIN A

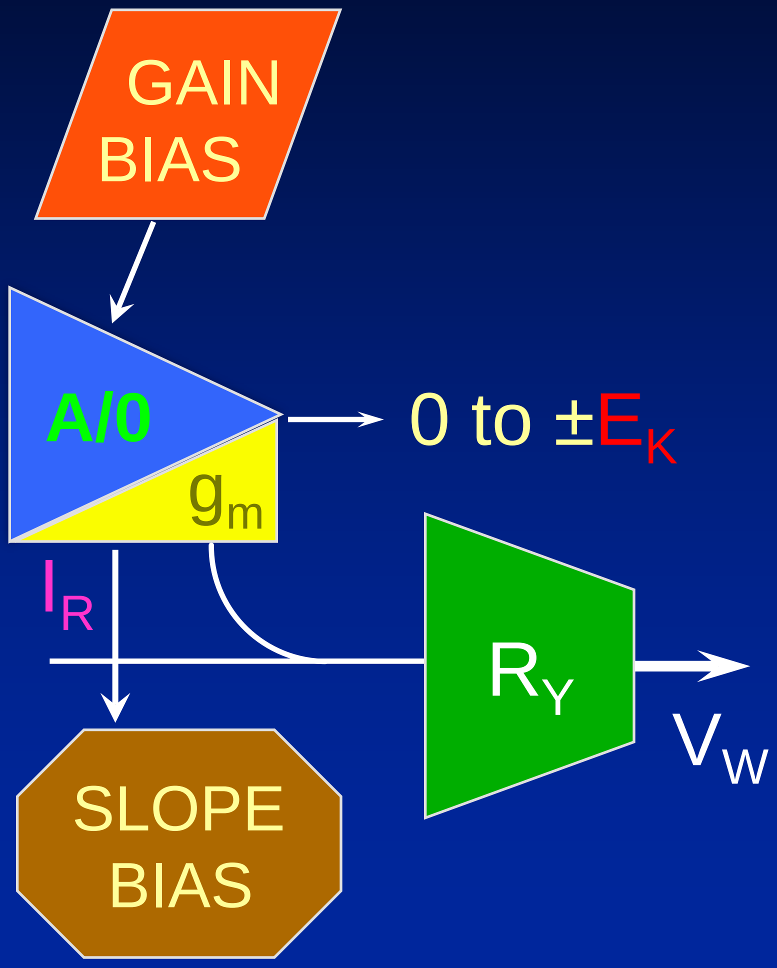


VOLTAGE REFERENCE SETS ACCURATE BIAS I_R

 BIW-10, Santa Fe, May 2010

ELIMINATION OF PTAT E_K

- OUTPUT CURRENT OF THE gm CELL CHANGES BY I_R FOR A CHANGE IN RATIO OF A AT THE AMPLIFIER'S INPUT
- CONVERTED TO A VOLTAGE USING A TRANSRESISTANCE OUTPUT STAGE, OF VALUE R_Y
- THUS, THE CHANGE IN V_W IS $I_R R_Y$ FOR EVERY FRACTION lgt(A) AT THE MAIN INPUT

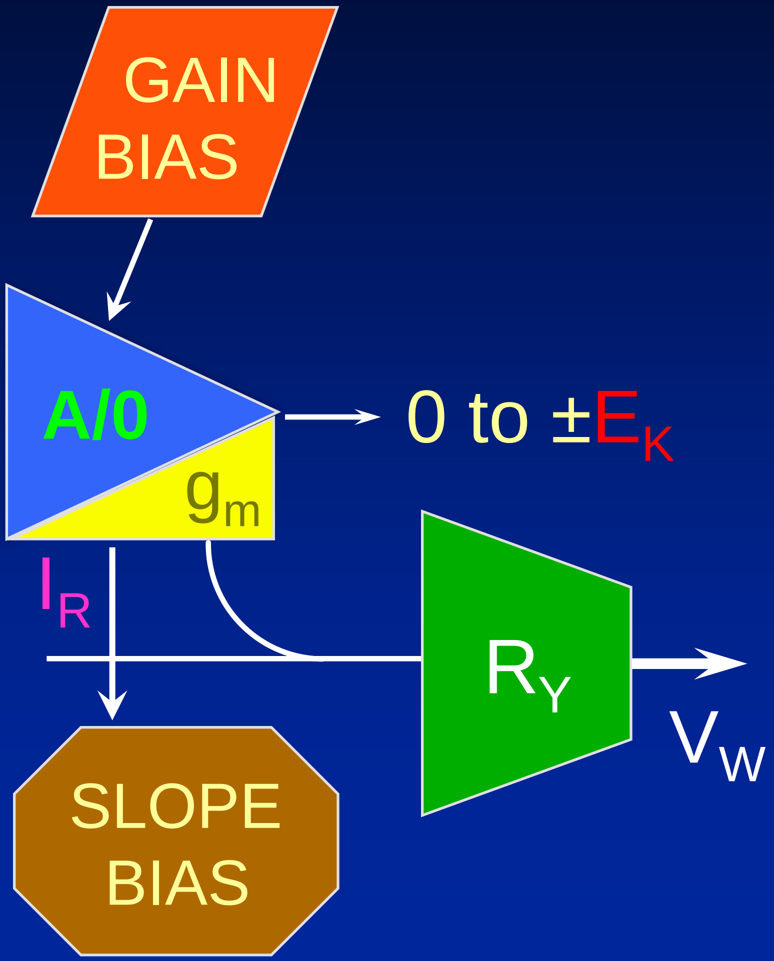


ELIMINATION OF PTAT E_K

- THE NEW SLOPE IS

$$V_Y = \frac{I_R R_Y}{\mathrm{lgt}\,(A)}$$

- THE DEPENDENCE ON A PTAT E_K HAS THUS BEEN ELIMINATED

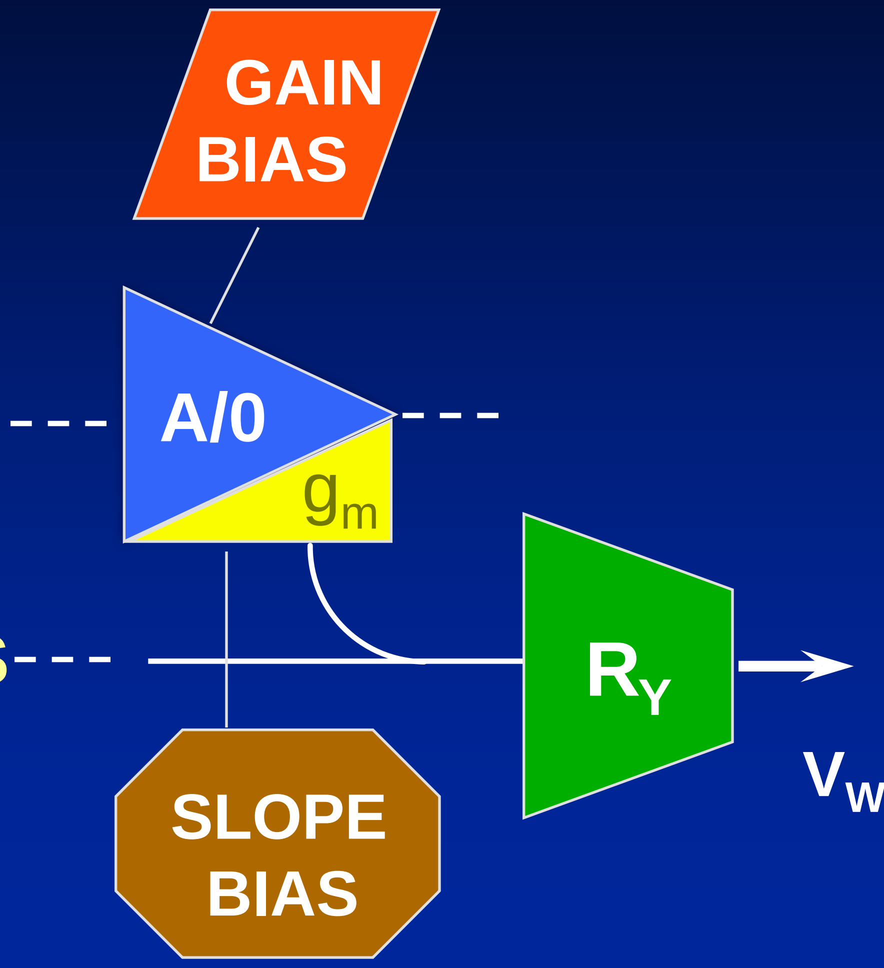


ELIMINATION OF PTAT E_K

- **THE NEW SLOPE IS**

$$V_Y = \frac{I_R R_Y}{\lg t\ (A)}$$

- **THE DEPENDENCE ON THE PTAT E_K HAS THUS BEEN ELIMINATED**



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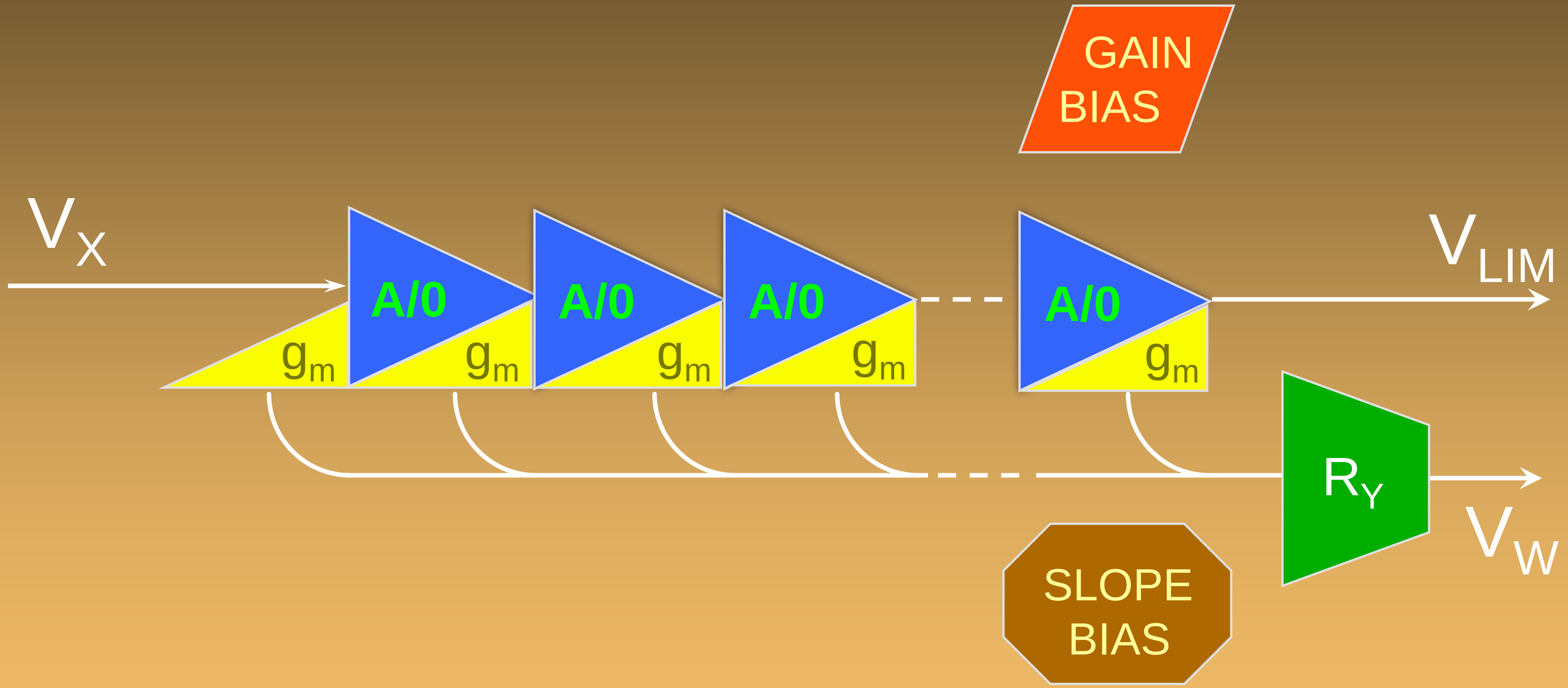
COMPLETE A/0 LOG-AMP



HOWEVER, THE INTERCEPT IS STILL DEPENDENT ON E_K

$$V_Z = \frac{E_K}{A^{N + 1/(A-1)}}$$

… and E_K in the 'natural' system comprising the log amp backbone is inherently PTAT.

The next few steps demonstrate how this can readily be compensated, to provide a stable intercept versus temperature.

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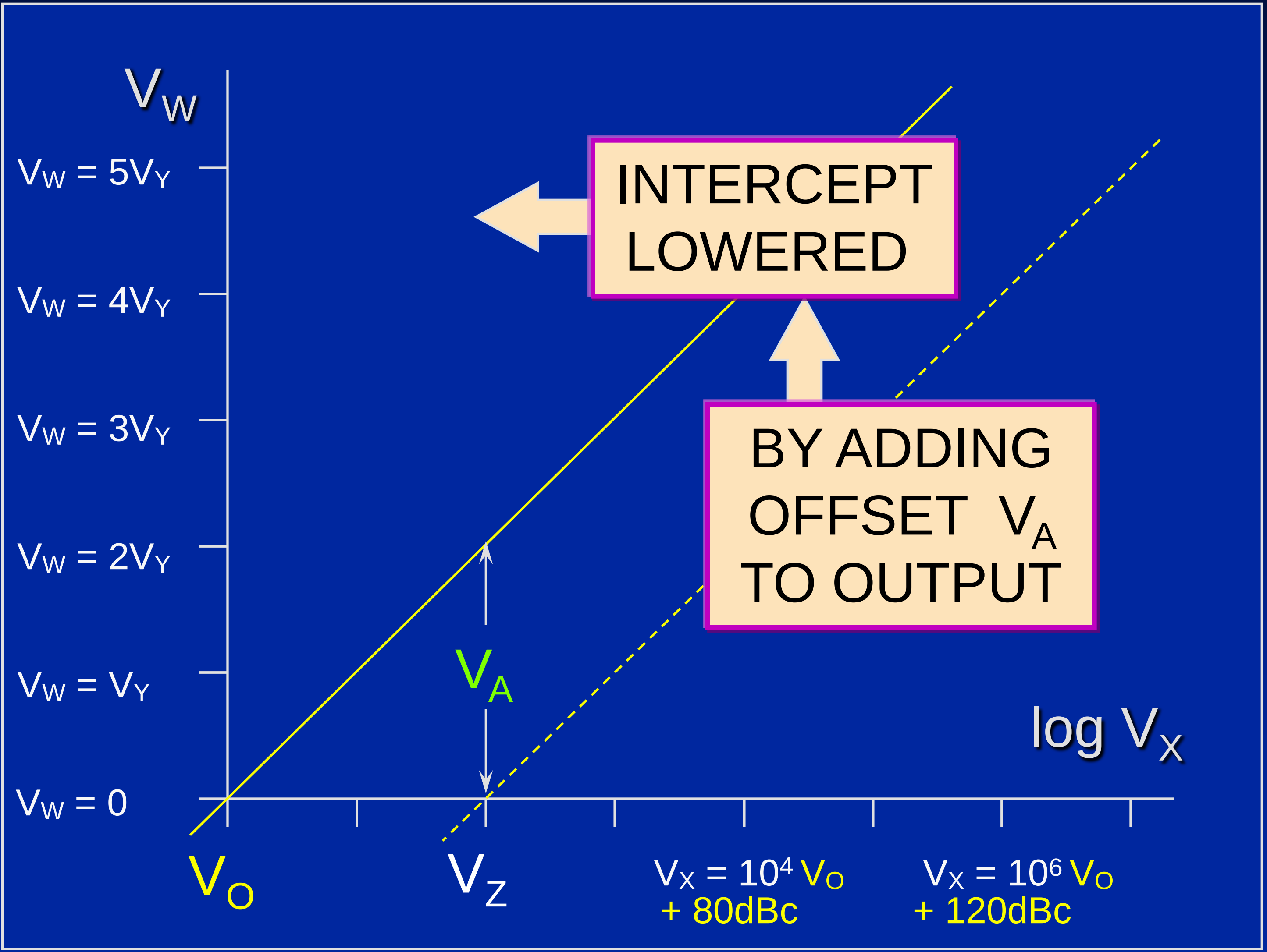
INTERCEPT MANIPULATION



INTERCEPT MANIPULATION

$$V_W = V_Y \log (V_X/V_Z)$$

$$= V_Y \log (V_X V_O / V_Z V_O)$$

$$= V_Y \log (V_X/V_O) + V_A$$

where $V_A = V_Y \log (V_O / V_Z)$

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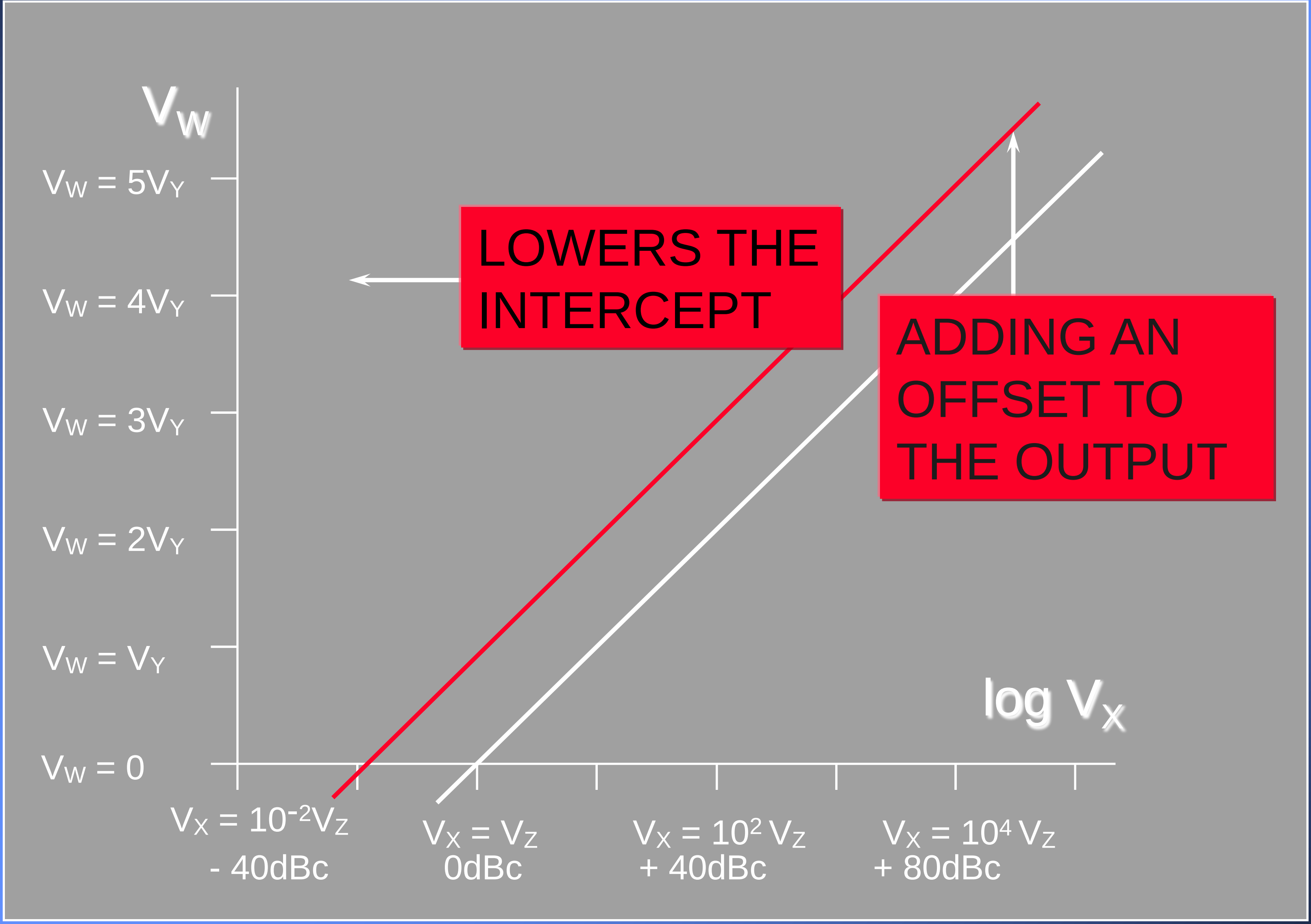
INTERCEPT POSITION IS EASILY ALTERED



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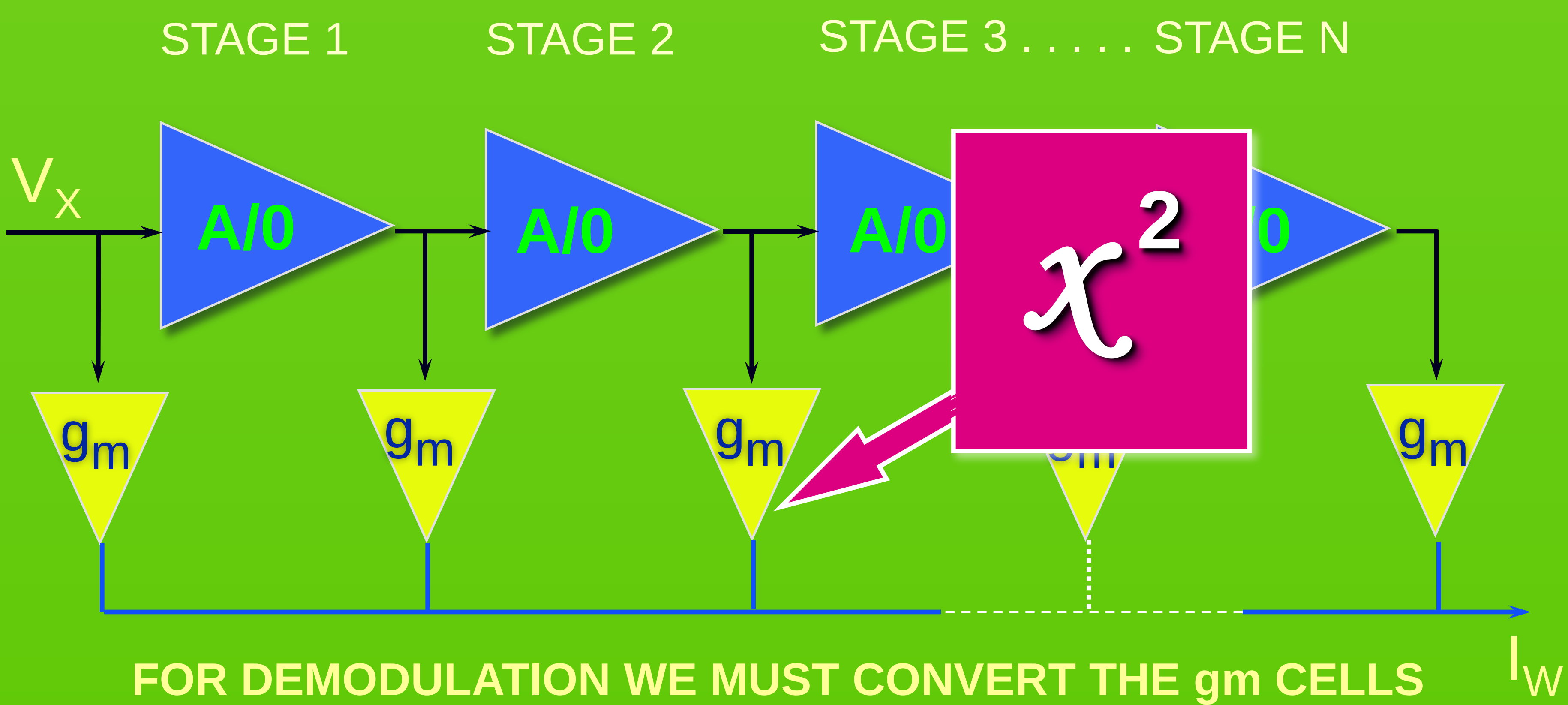
AUXILIARY g_m CELLS SUM IN CURRENT-MODE



FOR DEMODULATION WE MUST CONVERT THE gm CELLS INTO RECTIFIERS (DETECTORS) PREFERABLY FULL-WAVE

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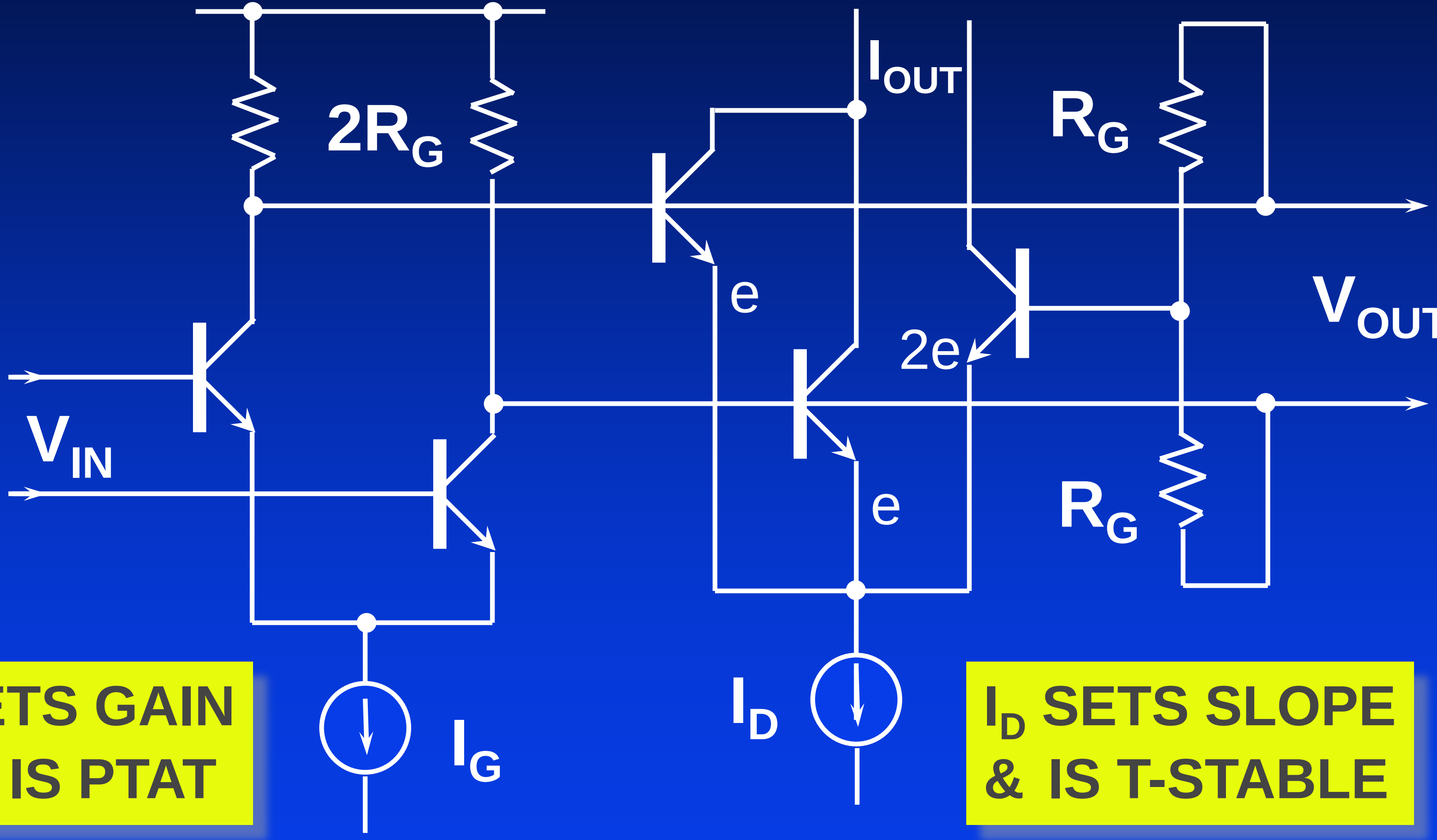
SQUARE-LAW DETECTOR



I_G SETS GAIN AND IS PTAT

I_D SETS SLOPE & IS T-STABLE

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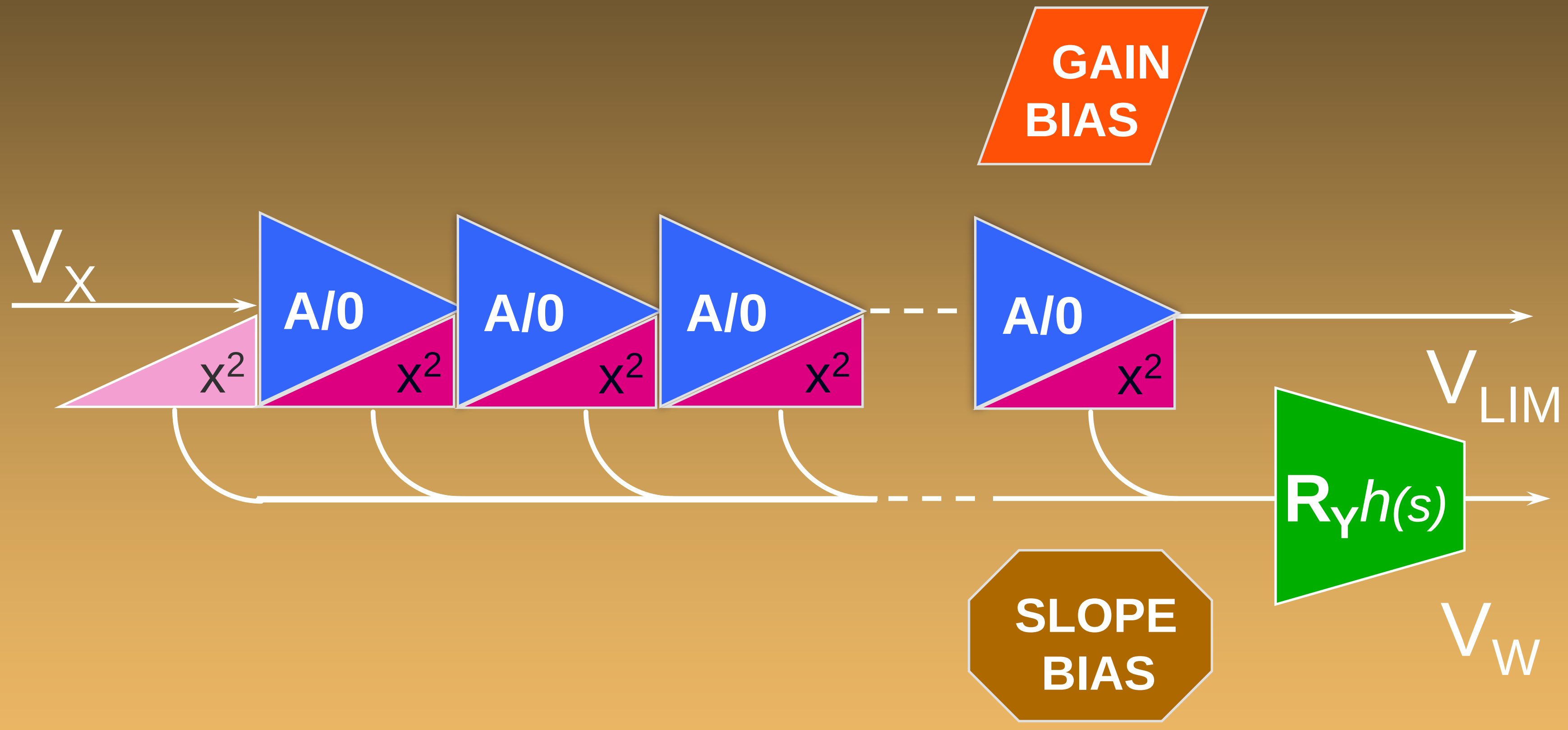
DEMODULATING LOG-AMP



THE OUTPUT I-V CONVERTER IS NOW ALSO A FILTER

APPROXIMATION RIPPLE

- THE LOG CONFORMANCE ERROR WHEN USING tanh GAIN CELLS NOW TAKES ON AN ESSENTIALLY SINUSOIDAL FORM. FOR A = 4 THE RIPPLE IS CLOSE TO ±0.2dB.
- SMALL ADJUSTMENTS ARE MADE TO THE WEIGHTING OF THE gm CELLS TO IMPROVE ACCURACY OF THE LOGARITHMIC FIT

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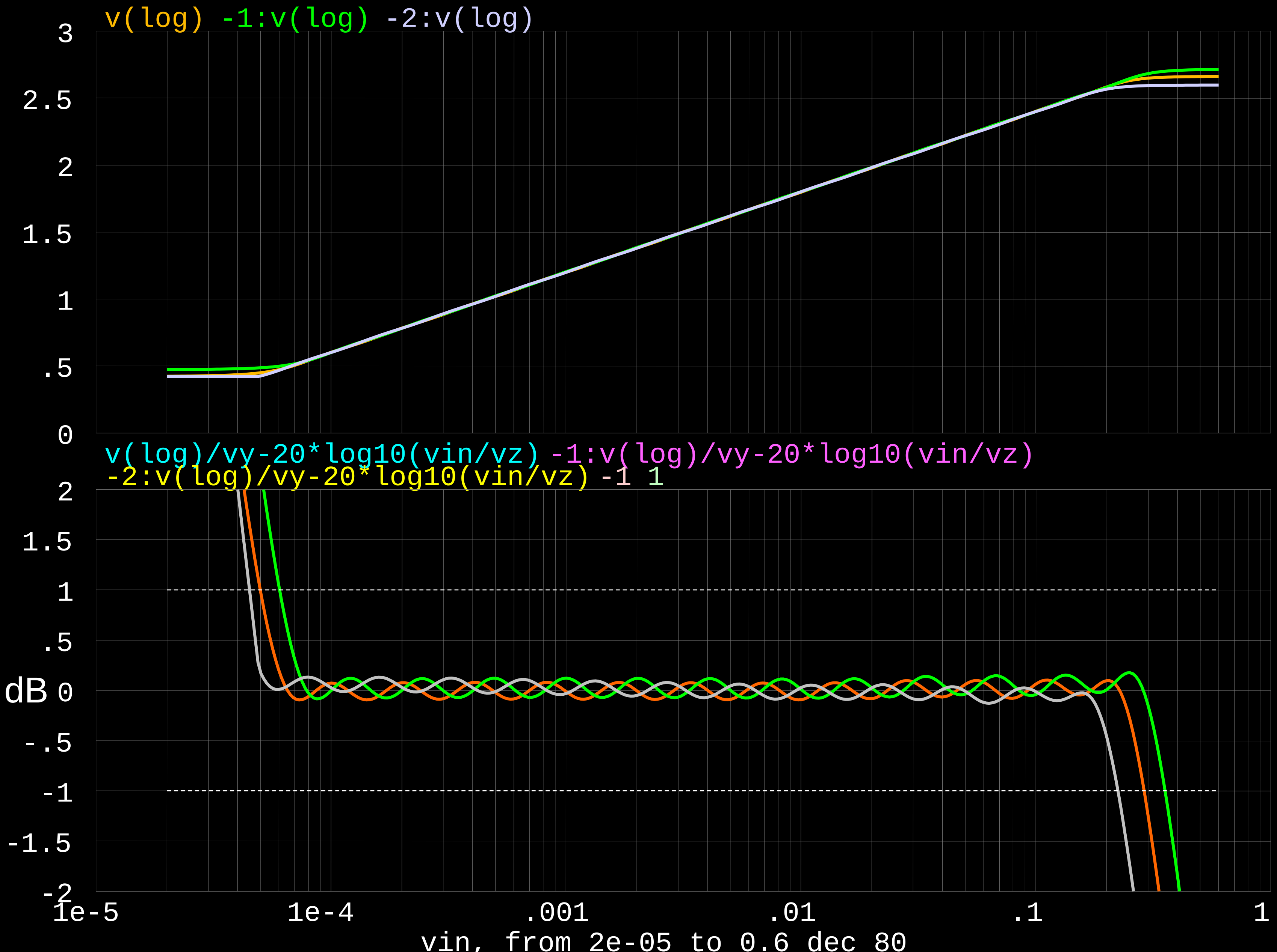

DYNAMIC RANGE ISSUES

RECALL THAT A BASIC REQUIREMENT OF A LOG-AMP IS THAT IT MUST HAVE VERY ZERO-SIGNAL GAIN, TO ACCURATELY RESPOND TO VERY SMALL INPUTS.

CHAINED A/0 AMPLIFIERS HAVE A GAIN OF A^N. FOR EXAMPLE, FOR A = 4 (12dB) & N=8, THE GAIN IS 65,500 (~96dB).

DYNAMIC RANGE ISSUES

FOR A MODEST OVERALL BANDWIDTH OF 3.2 GHz, AND THIS GAIN OF 65,000, THE *GAIN-BANDWIDTH PRODUCT* IS IS AN ASTRONOMICAL 208,000 GHz !

DYNAMIC RANGE ISSUES

THE DYNAMIC RANGE OF A LOG-AMP IS LIMITED BY ITS FIRST-STAGE NOISE (say, 1nV/√ Hz) & BY THE PEAK INPUT AT WHICH IT REMAINS LOGARITHMIC, ROUGHLY ± $2E_K$. THIS IS ABOUT 74dB.

SOLUTIONS

ONE SOLUTION IS TO LOWER THE BANDWIDTH, BUT THIS MAY NOT BE PERMISSIBLE. ANOTHER IS TO LOWER THE NOISE; THAT'S HARD.

A BETTER SOLUTION IS TO RAISE TOP END OF THE DYNAMIC RANGE BY USING AN AUXILIARY LOG-AMP

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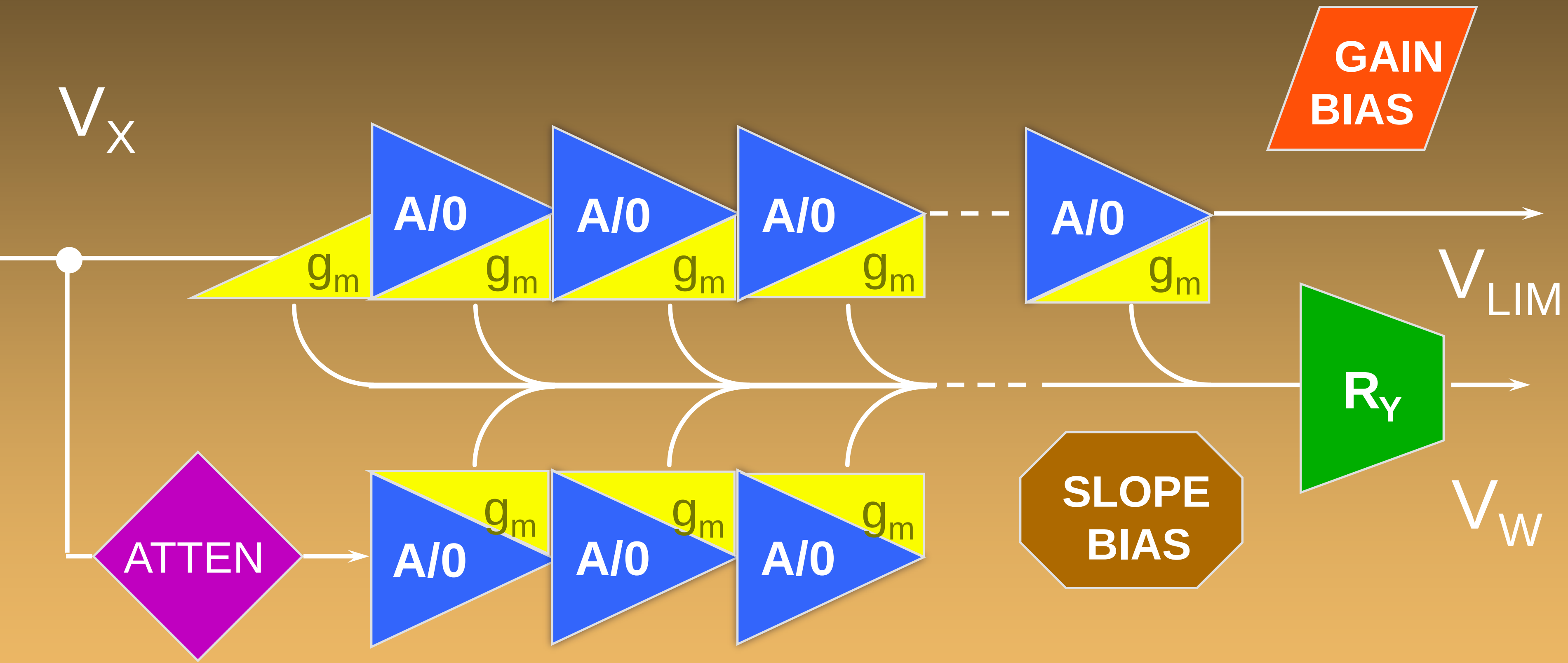
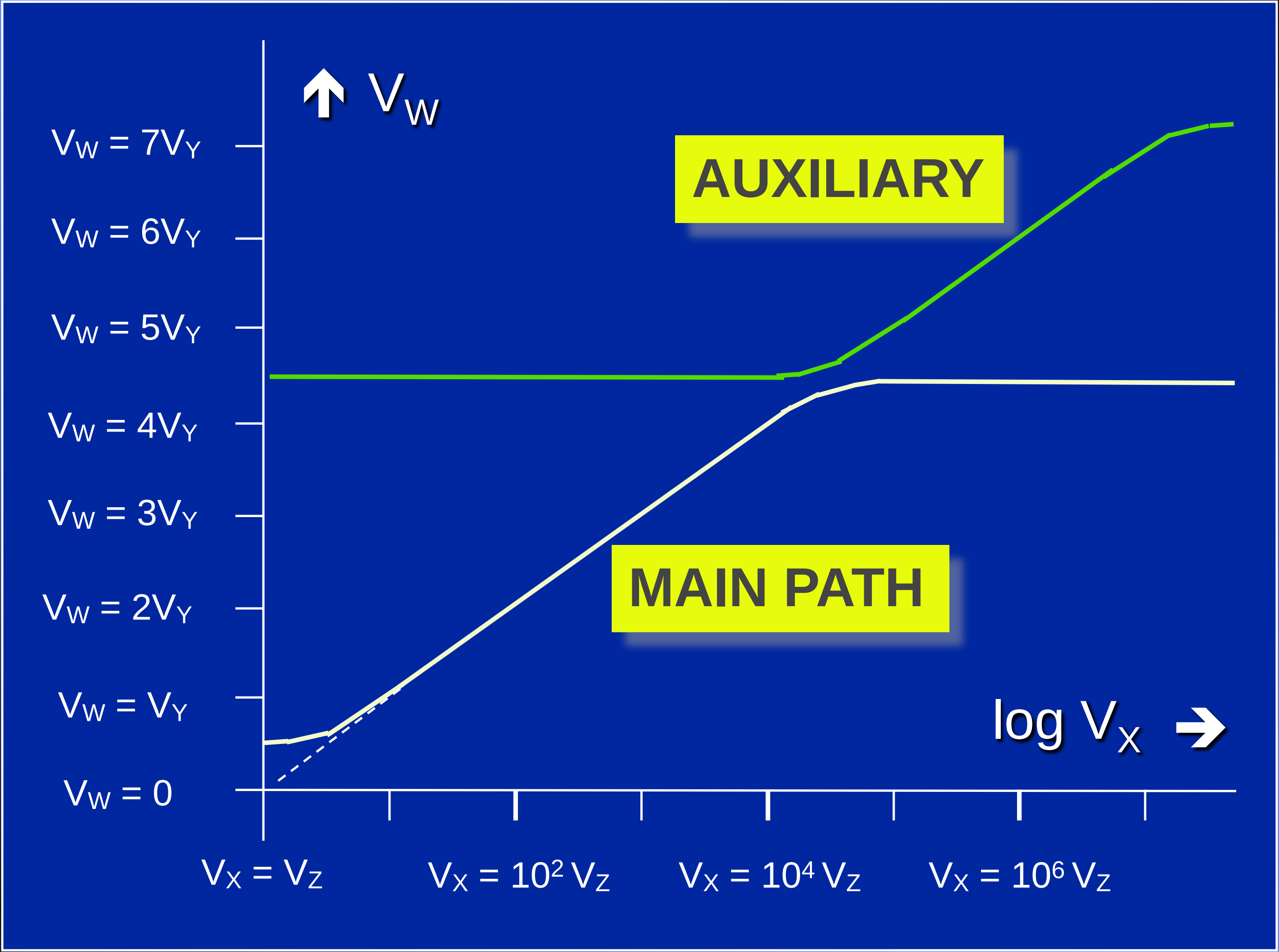
WIDE RANGE LOG-AMP



ATTENUATION RATIO MUST BE A^K WITH K INTEGER

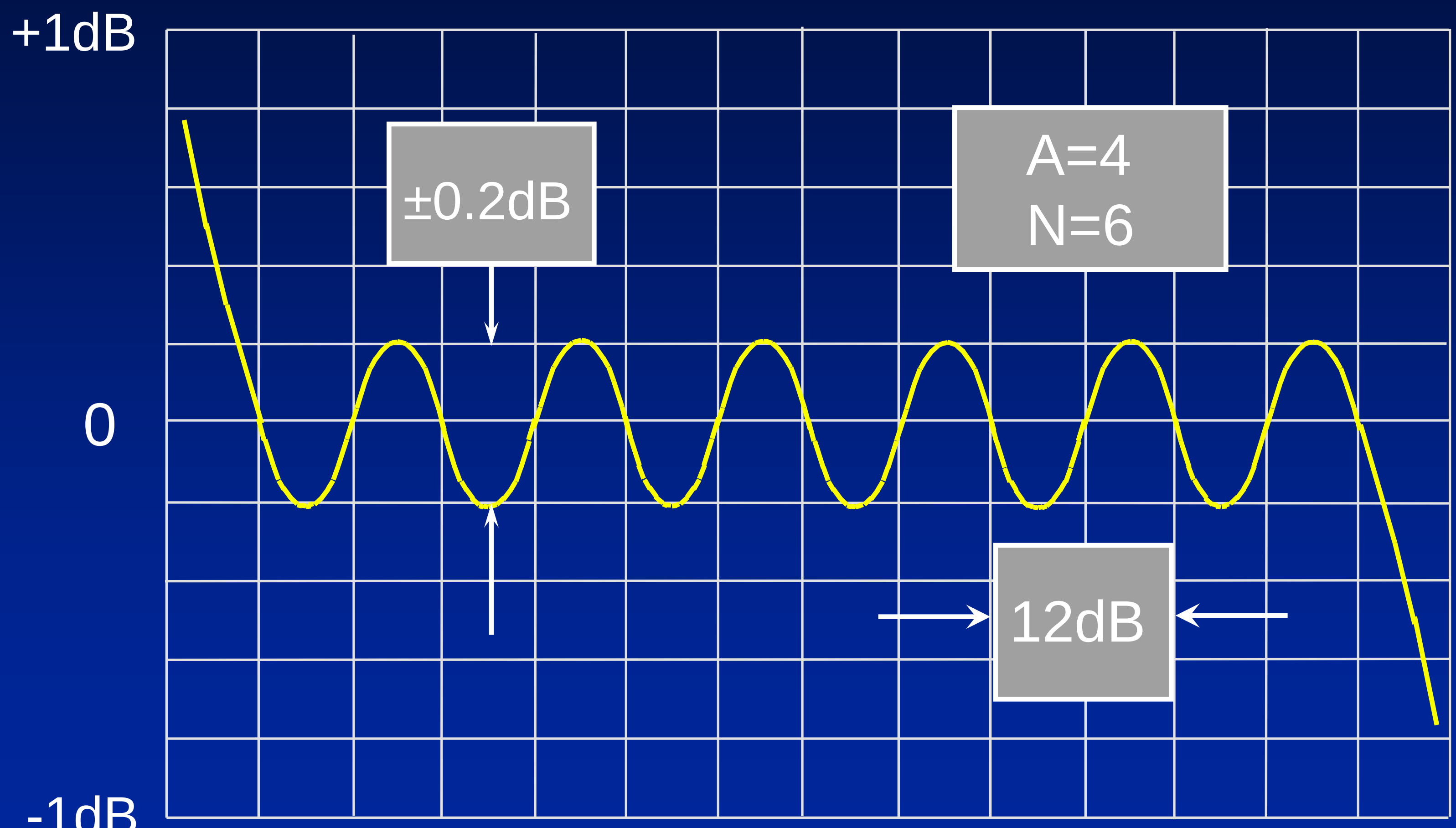
WIDE RANGE LOG-AMP



WIDE RANGE LOG-AMP

USING THIS TECHNIQUE, A DYNAMIC RANGE OF >100dB CAN BE ACHIEVED, EVEN AT WIDE BANDWIDTHS. WITH 1nV/√Hz AND Δf = 400MHz, THE NOISE VOLTAGE IS 20μV RMS; THE LARGEST INPUT MAY BE LIMITED ONLY BY BV_{CBO}, SAY, 4V, ABOUT 2.8V RMS. THIS IS A DYNAMIC RANGE OF 103dB.

APPROXIMATION ERROR FOR NON-IDEAL (tanh) CELLS (DC)



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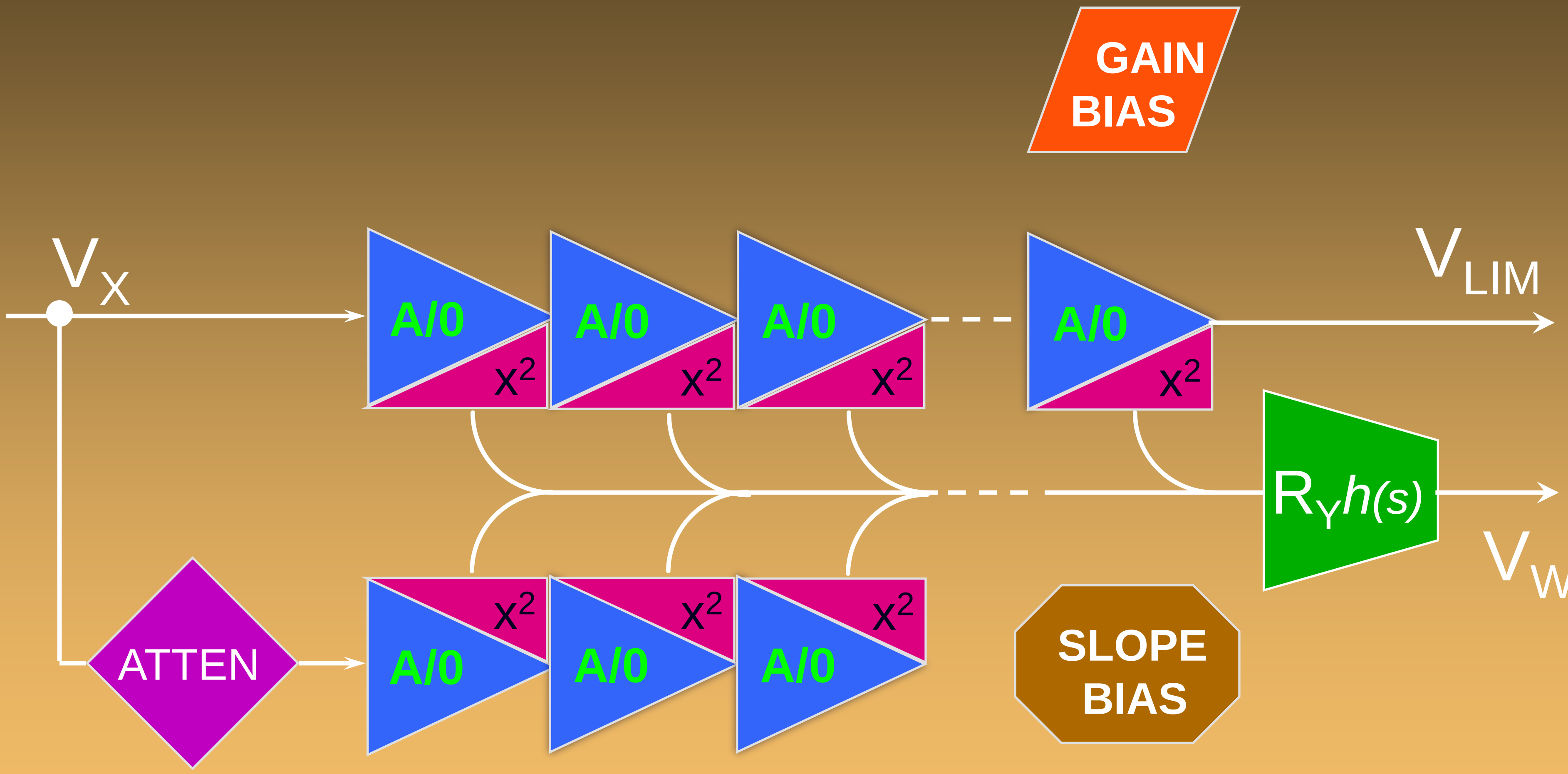
WIDE RANGE LOG-AMP



THE OUTPUT I-V CONVERTER IS NOW ALSO A FILTER

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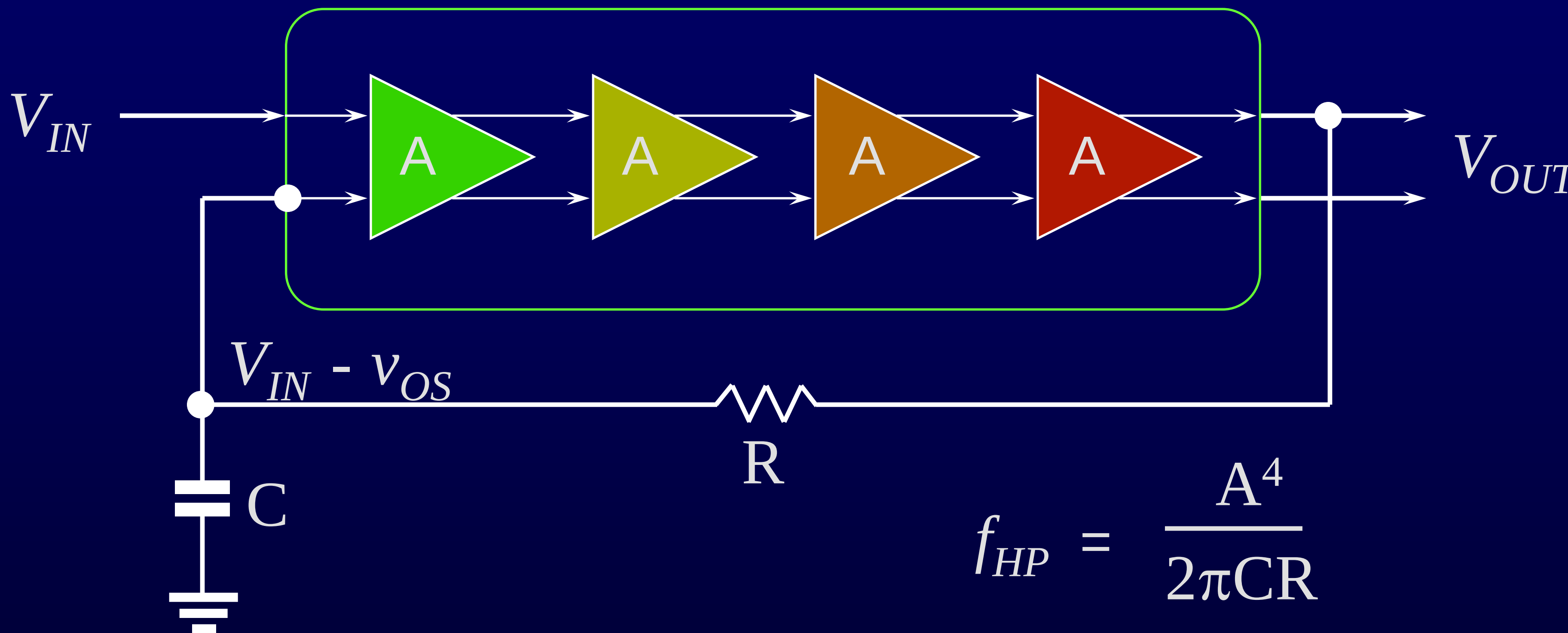
REMOVING DC OFFSET



$$f_{HP} = \frac{A^4}{2\pi CR}$$

To avoid breaking the chain, while at the same time cope with DC offsets a global feedback path can be used to stabilize the operating point (null the offset)

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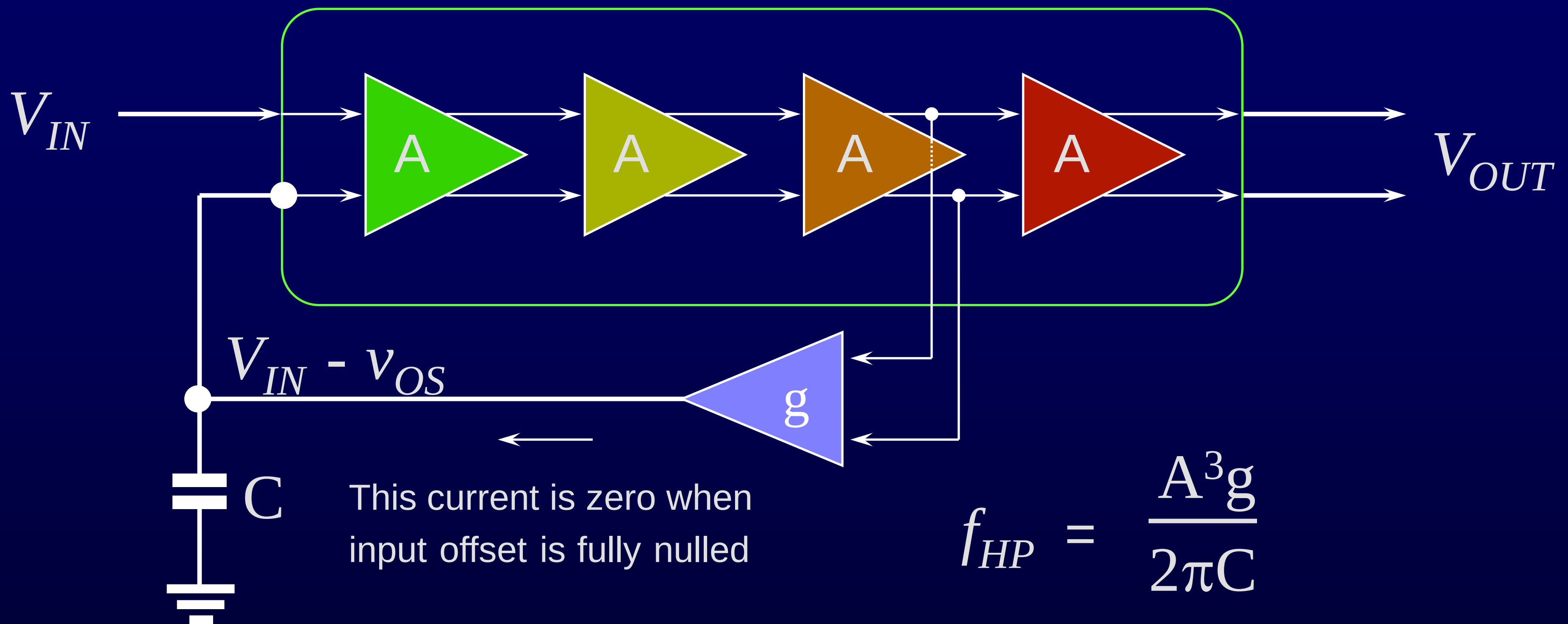
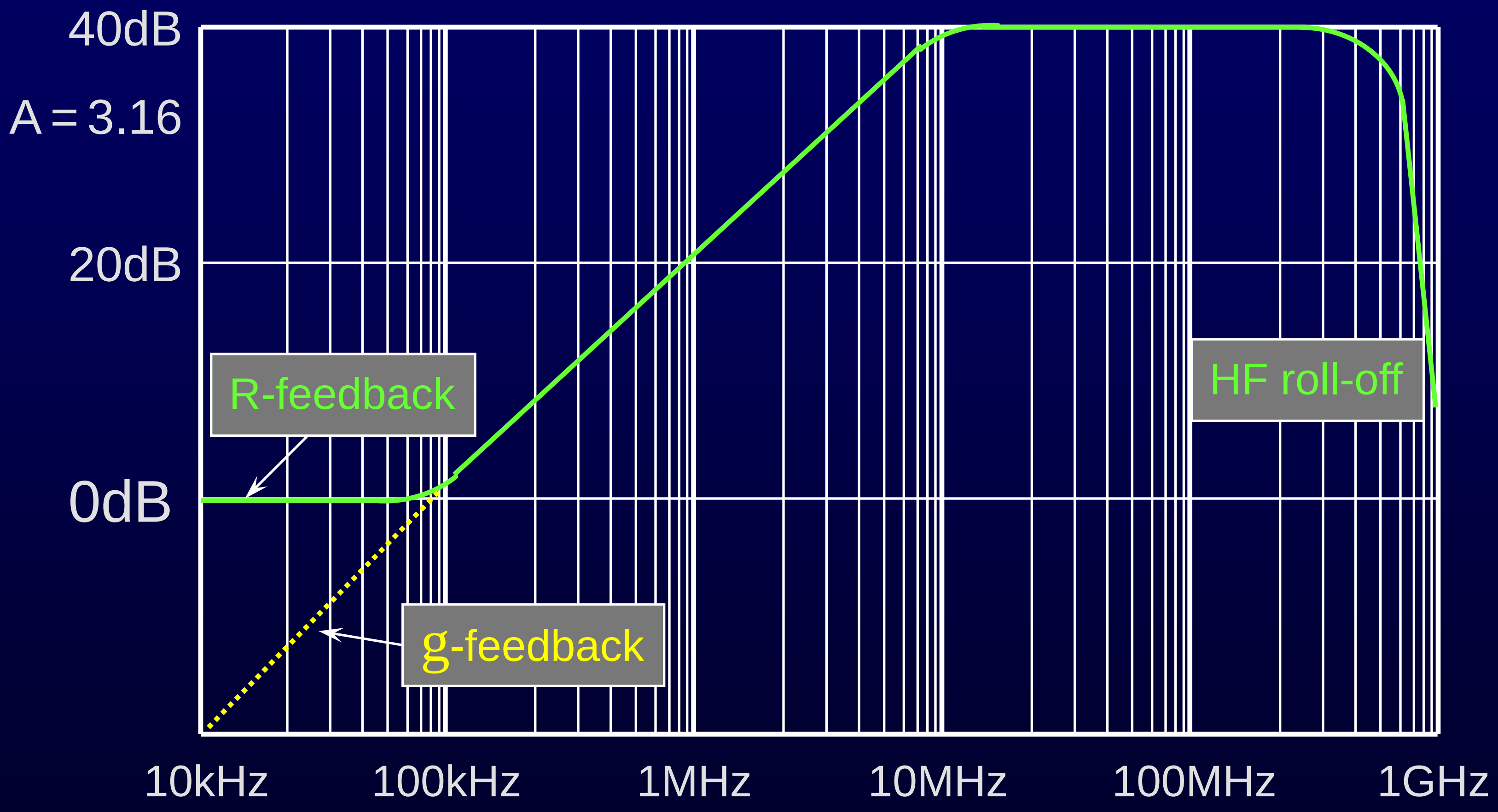
REMOVING DC OFFSET



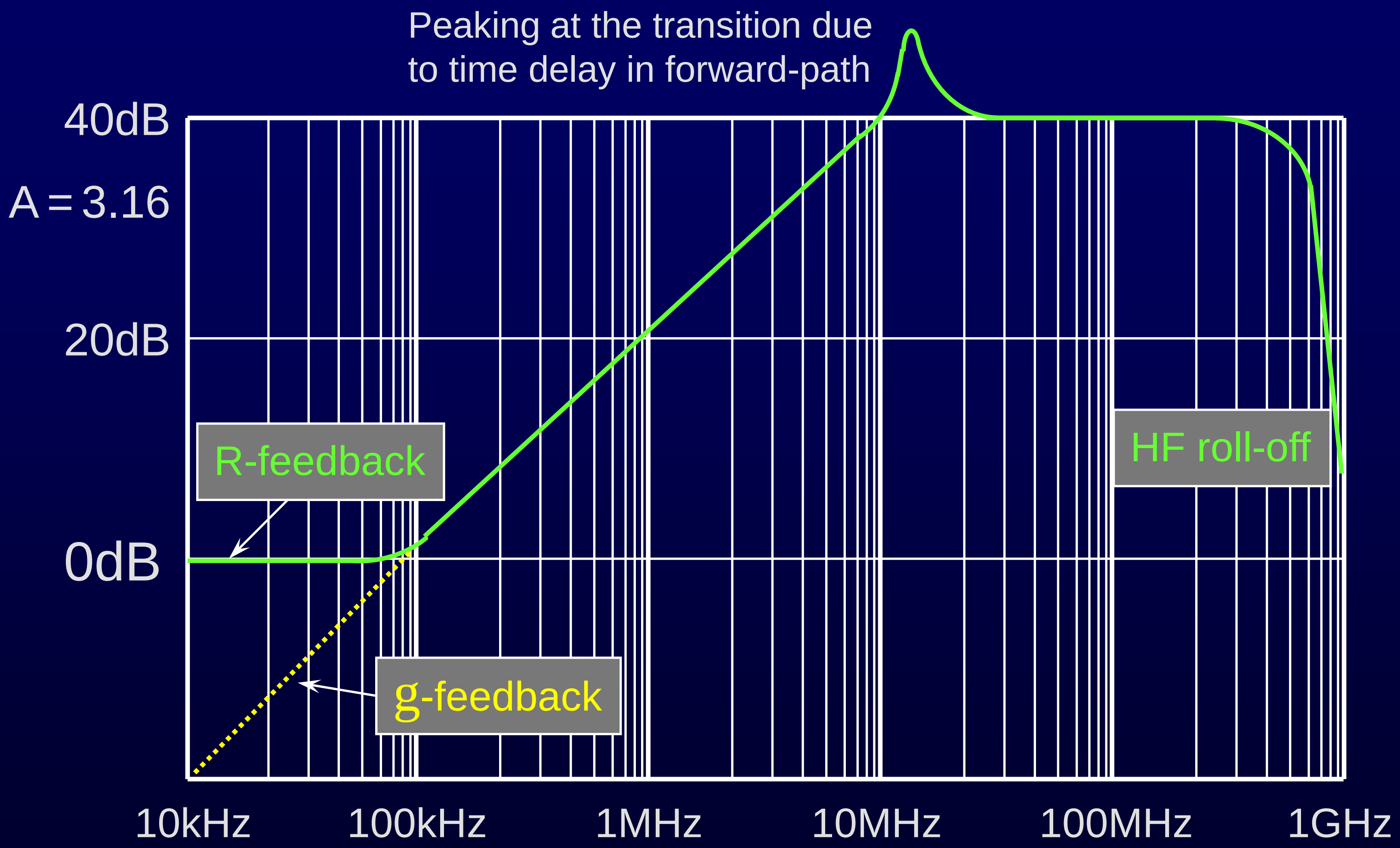
$$f_{HP} = \frac{A^3 g}{2\pi C}$$

This method is often useful in coping with DC offset since it forms an integrator that fully nulls the error.

REMOVING DC OFFSET



REMOVING DC OFFSET



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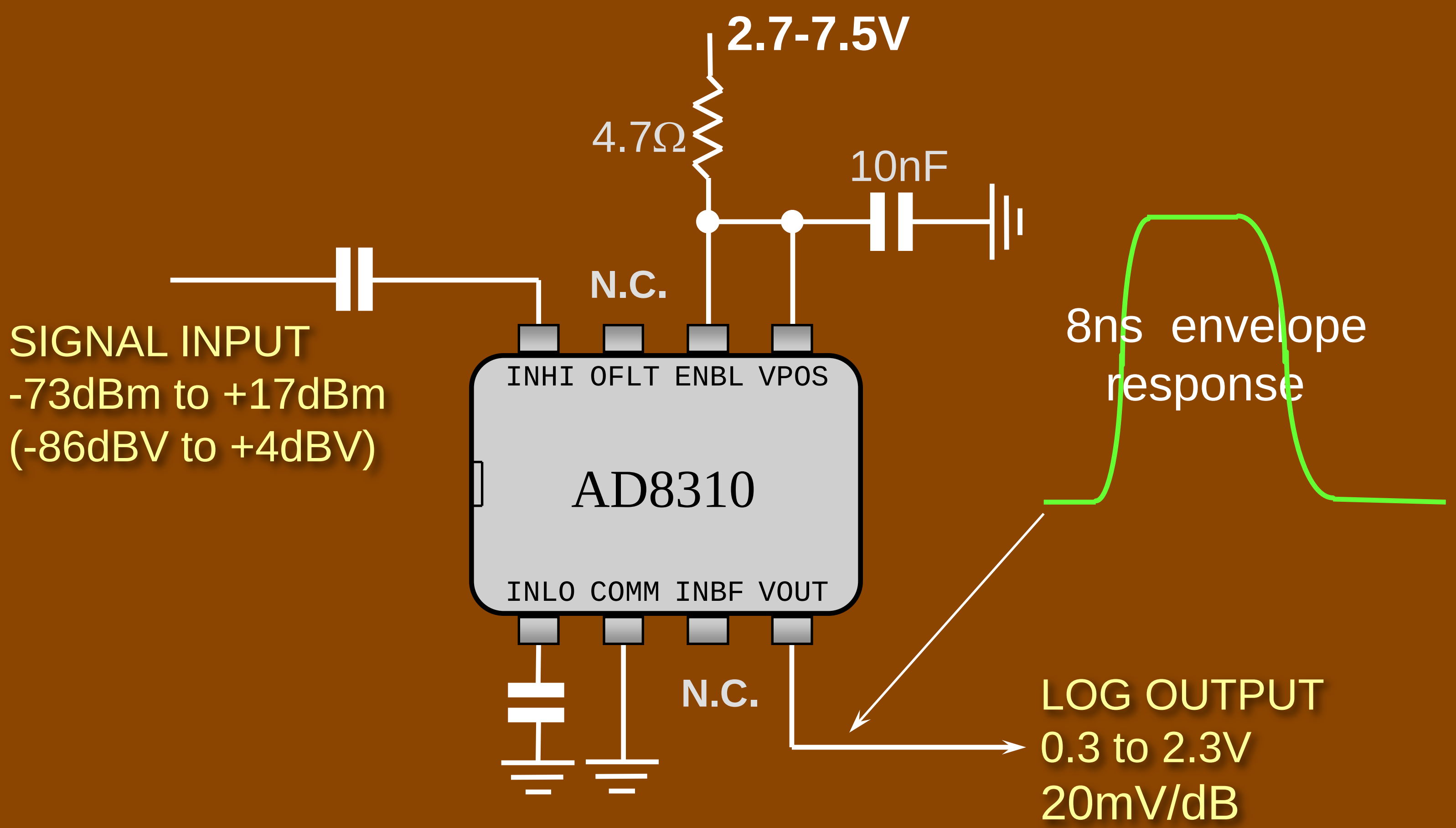
AD8310: Fast, Lo-Z Voltage Out



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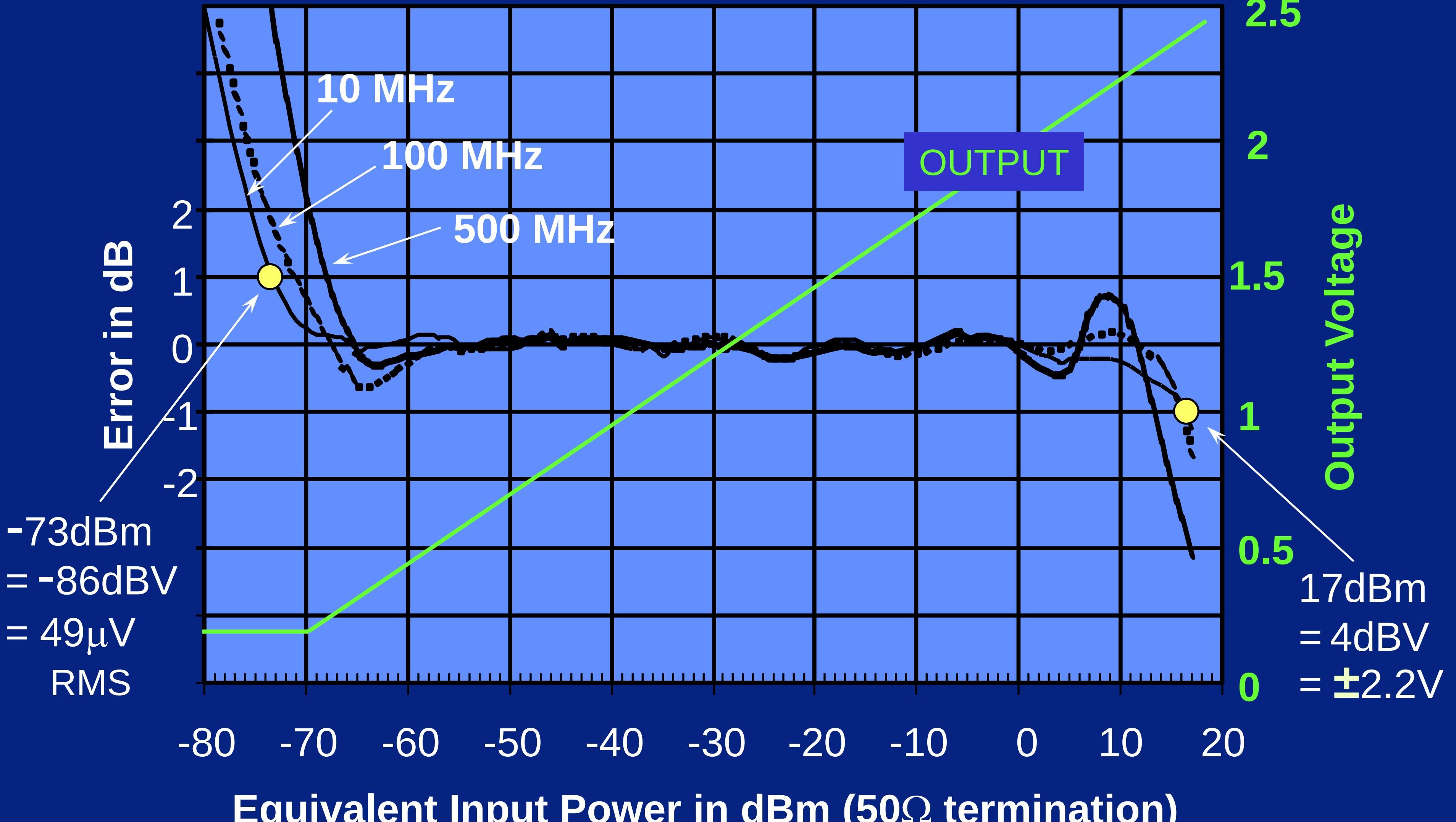
Logarithmic Conformance of AD8307 (from Data Sheet)



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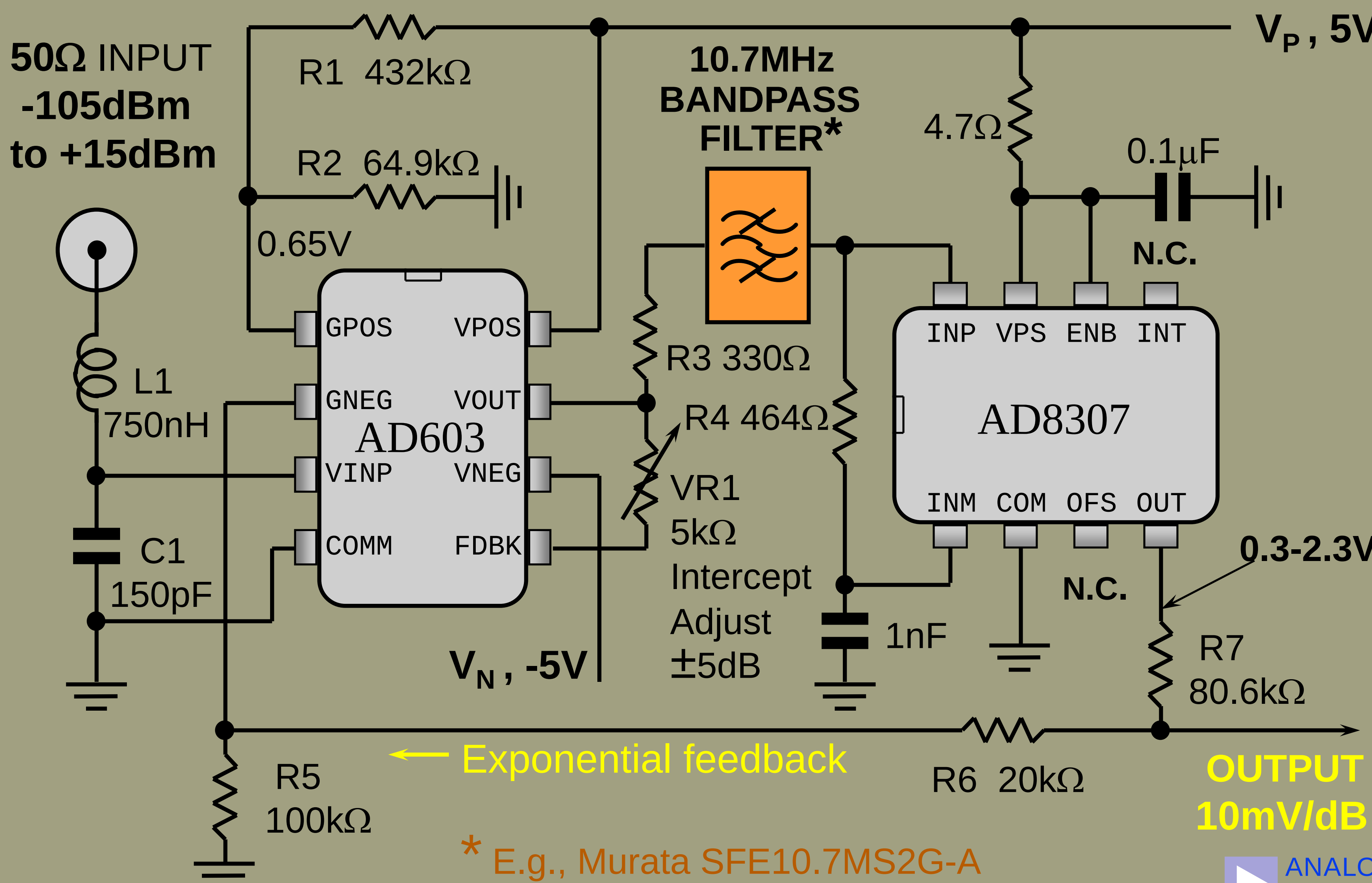
120+ dB MEASUREMENT SYSTEM



* E.g., Murata SFE10.7MS2G-A

ANALOG DEVICES

EFFECT OF WAVEFORM

- THESE ANALYSES HAVE BEEN FOR DC INPUTS ONLY
- AN AMPLITUDE-SYMMETRIC PULSE OR SQUAREWAVE INPUT WOULD PRODUCE THE SAME RESULT
- FOR STANDARD WAVEFORMS, IT IS EASY TO CALCULATE THE EFFECT ON THE CALIBRATION

EFFECT OF WAVEFORM

- WAVEFORM HAS NO EFFECT ON THE SLOPE CALIBRATION, V_Y
- IT AFFECTS ONLY THE INTERCEPT

EFFECT OF WAVEFORM

- FOR A SINUSOIDAL INPUT $E_S \sin\omega t$ THE OUTPUT WILL BE THE SAME AS THAT FOR A DC INPUT OF $E_S/2$
- THAT IS, INTERCEPT VALUE FOR SINE IS EFFECTIVELY DOUBLED

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RF Power Detectors

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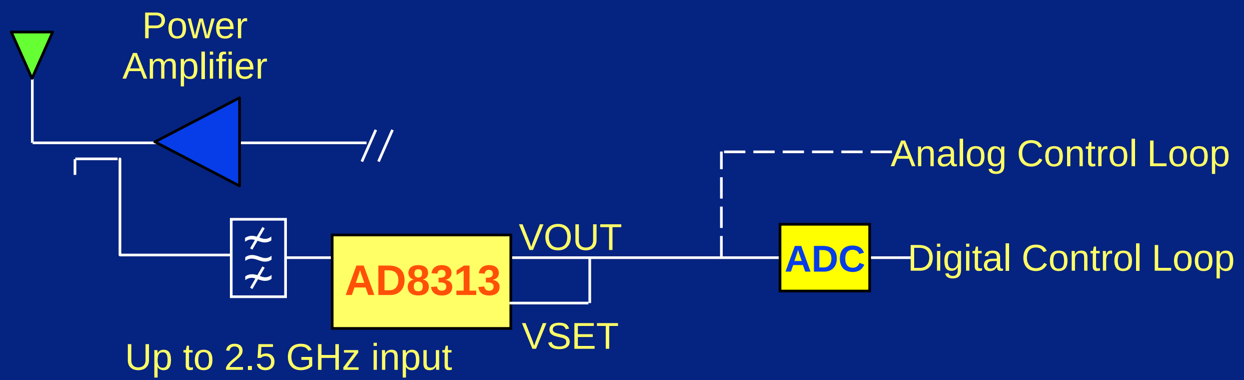
Direct RF Detection using AD8313

ANALOG DEVICES



- Direct RF Detection to >2.5 GHz
- 70 dB total dynamic range
- $\pm$ 1dB accuracy over central 62dB
- 8 pin Micro-SOIC
- Provides RSSI at antenna frequency
- Released August 1998

ANALOG DEVICES

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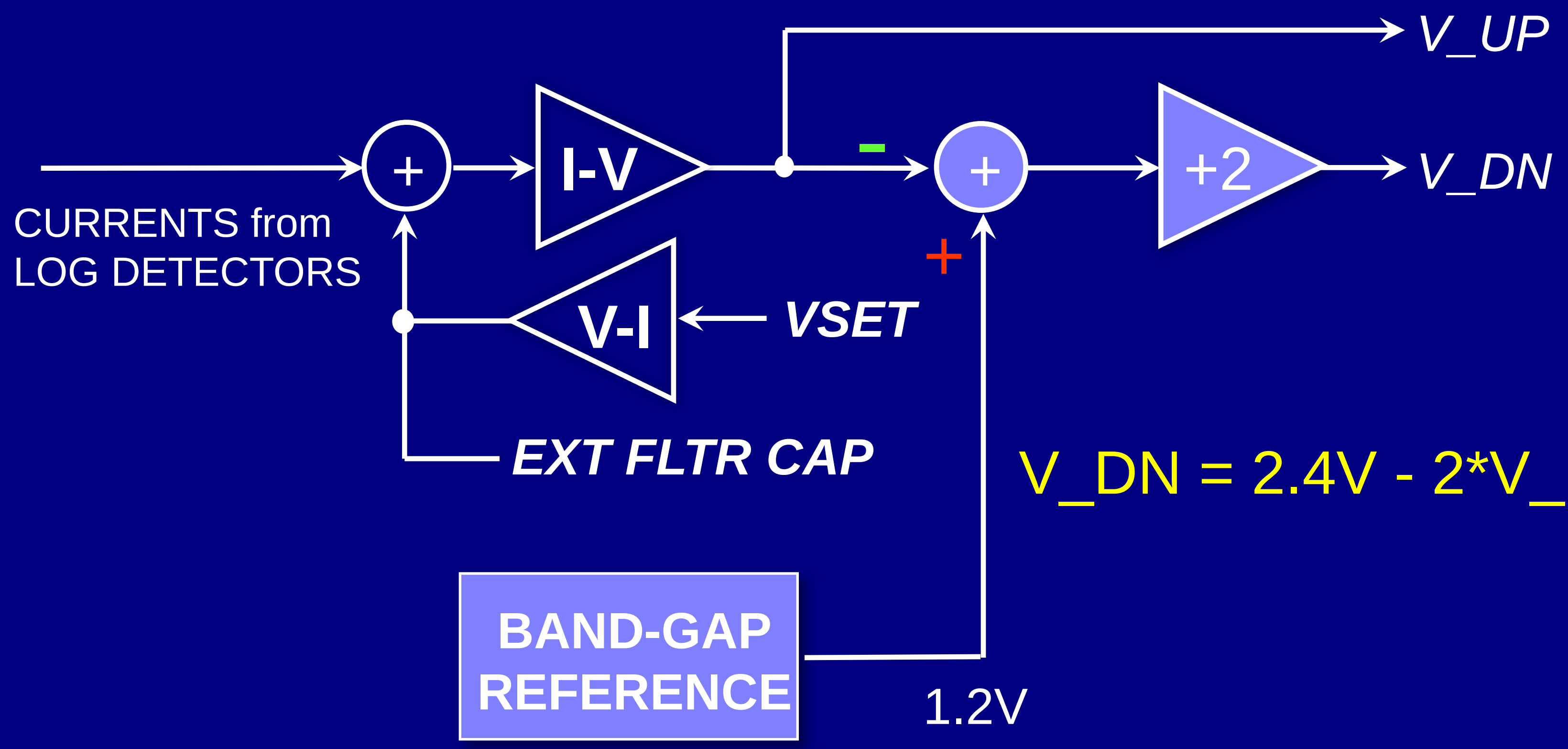
AD8314 Low Cost 40dB Log Amp



V_DN = 2.4V - 2*V_UP

Output Interface of AD8314

ANALOG DEVICES

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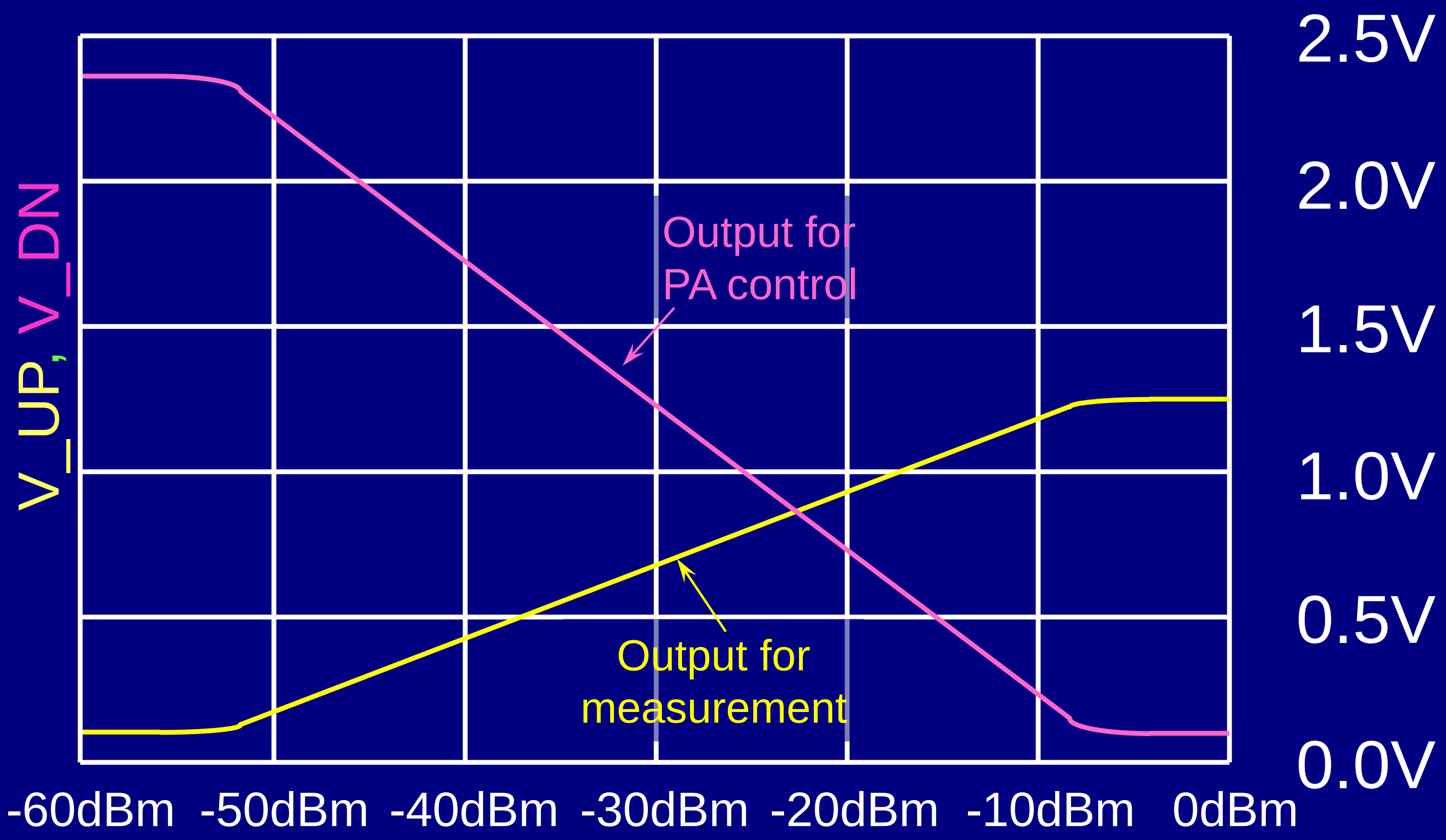
AD8314 40dB Log-Amp is a Detector & Controller



ANALOG DEVICES

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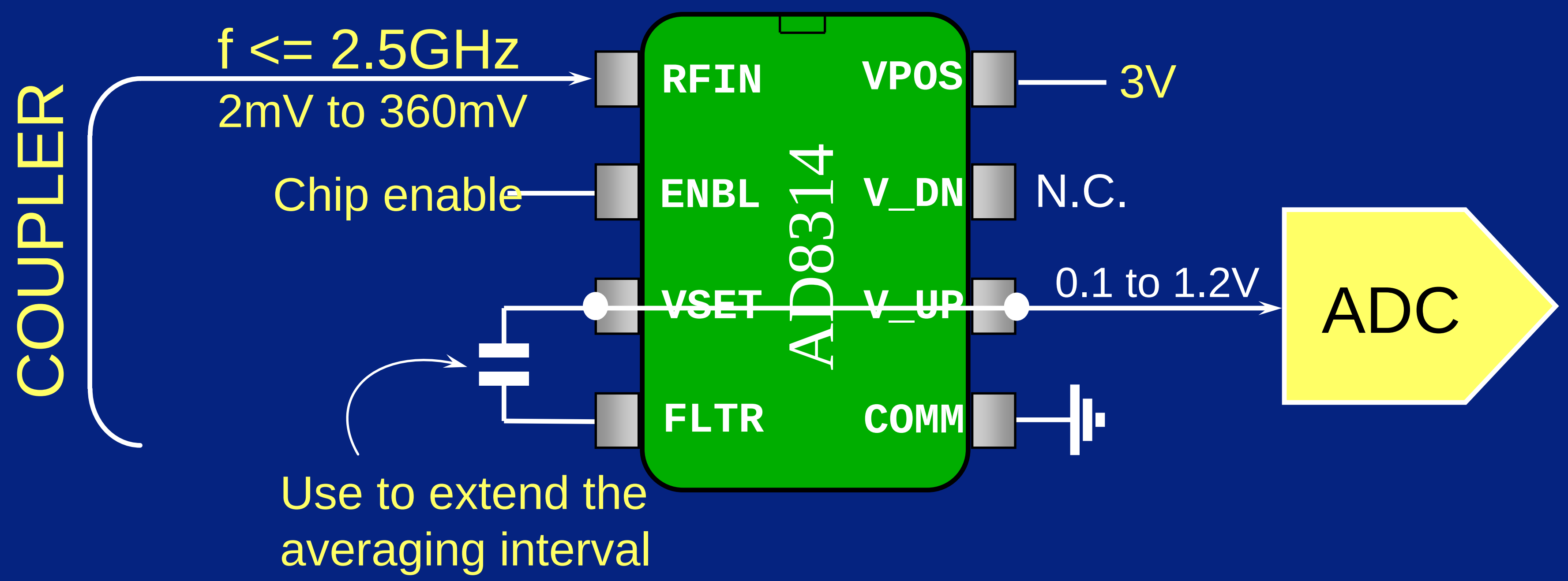
AD8314 in Measurement Mode



Output is 0.5 log_{10} (V_{IN} in mV)

ANALOG DEVICES

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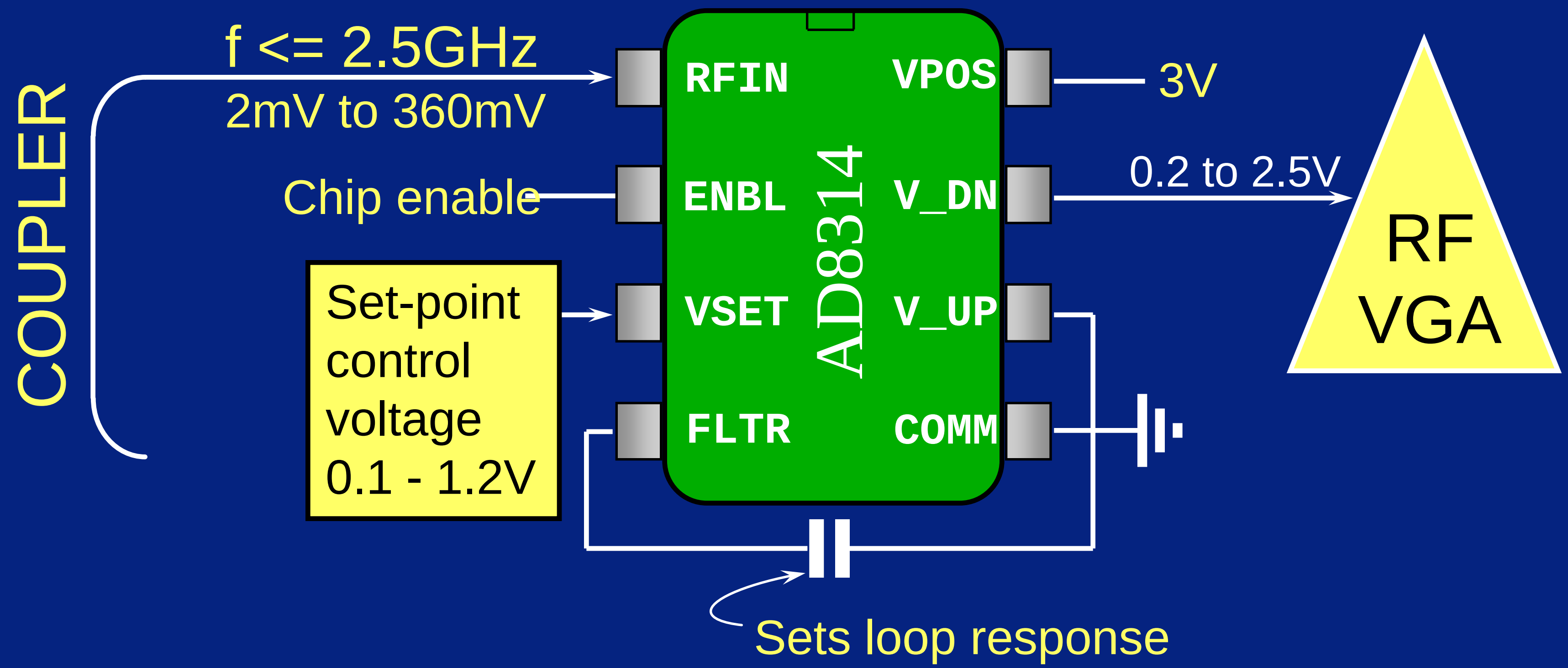
AD8314 in Controller Mode



V_UP not used as an output in this mode but is needed to set the loop bandwidth

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One-chip Network Analyser

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AD8302: A GAIN/PHASE DETECTOR

- A Network Analyser on a Chip! - Almost!

$$V_{GAIN} = V_G \, log \, (V_A/V_B) \qquad V_G = 30mV/dB$$

$$V_{PHS} = V_P \, (\phi_1 - \phi_2) \qquad V_P = 10mV/deg$$

- Operates from LF to >3 GHz

Applications

- Power Amplifier Phase/Gain Control
  .... independent of actual power level
- Monitoring of System Gain/Loss (e.g. Return Loss)
- System Diagnostics
- Linear Phase Demodulator

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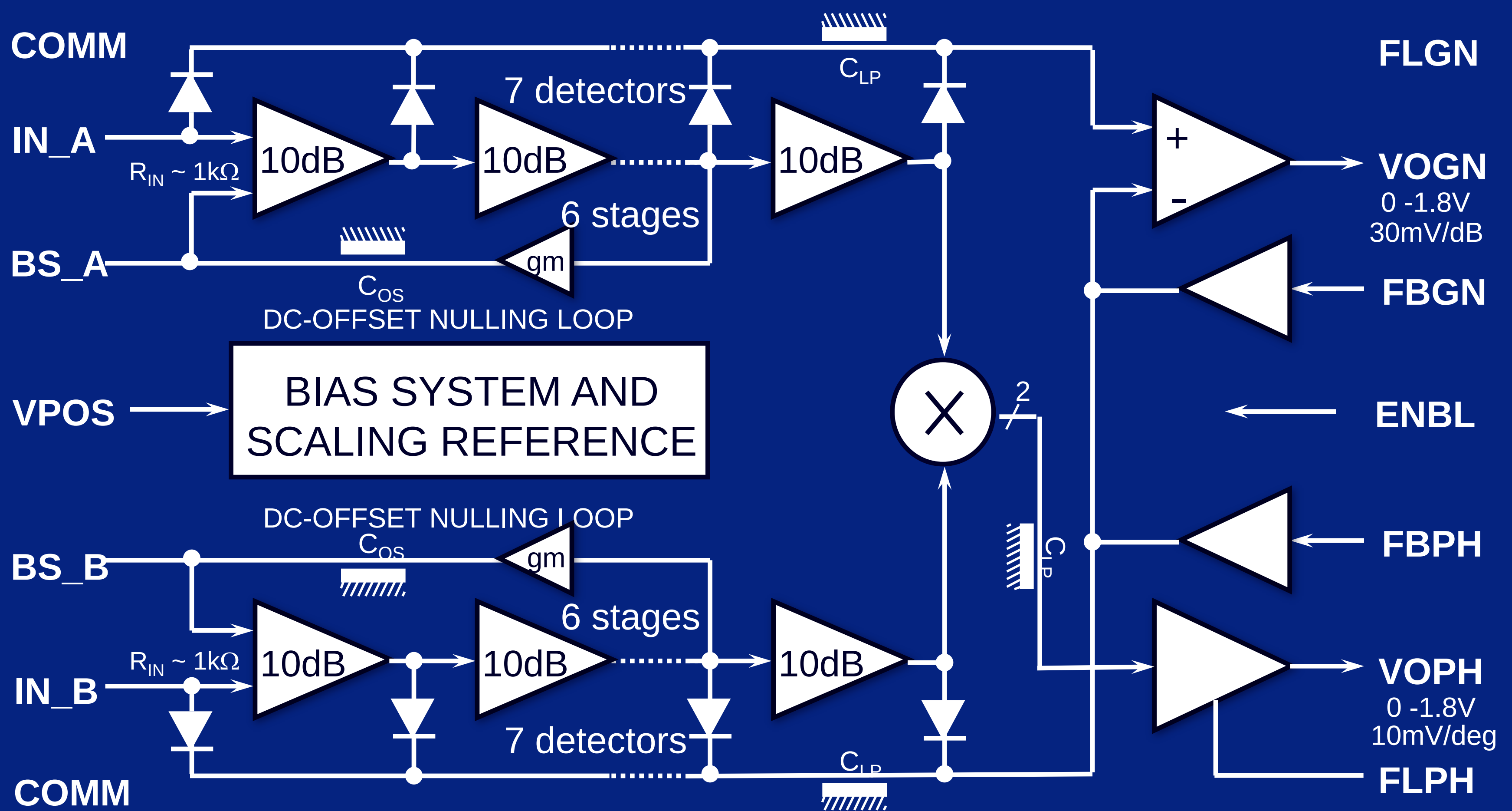
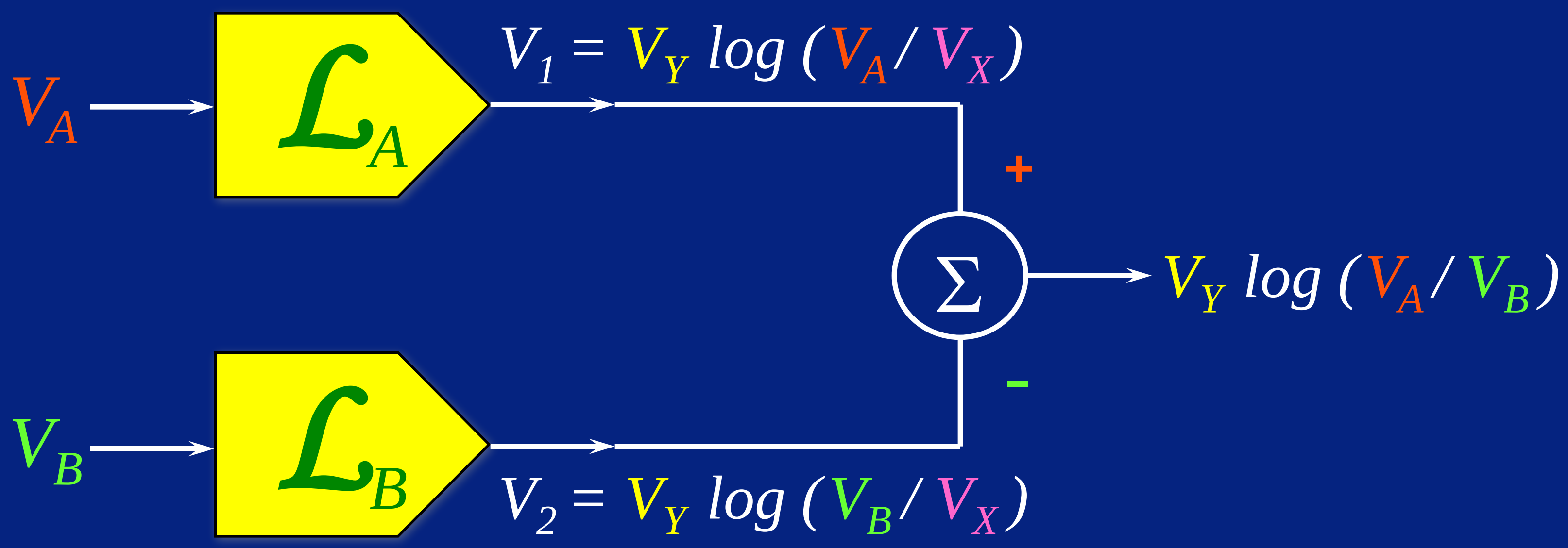
COMPLETE GAIN/PHASE DETECTOR



Gain/Phase Detector: General Scheme

TRUE GAIN MEASUREMENT



By subtracting the output of the B-channel log-amp from that of the A-channel log-amp, the intercept V_X is eliminated and the resulting difference is a measure of the RATIO of V_A / V_B

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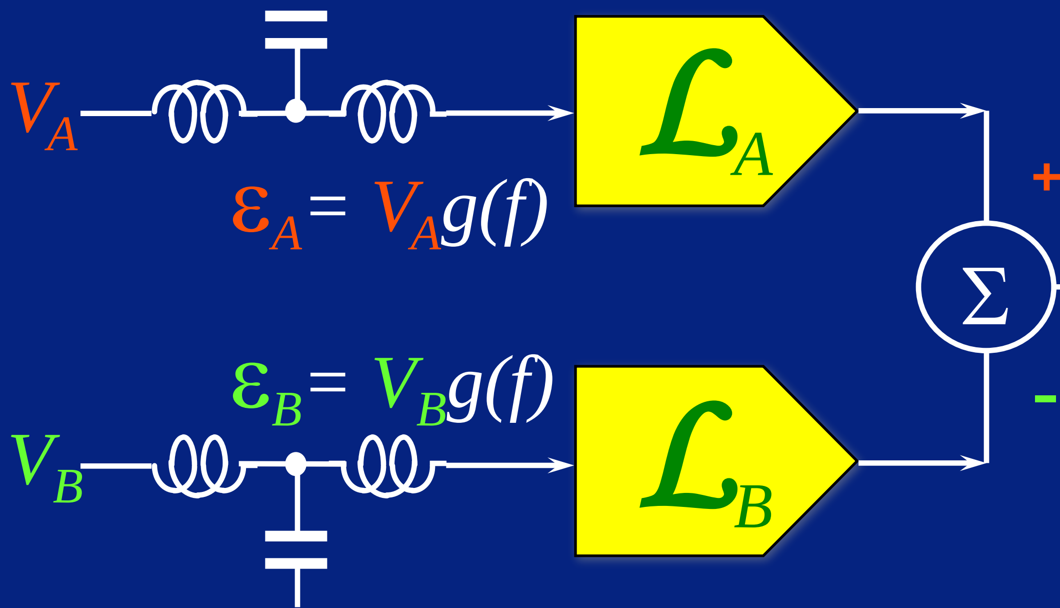
CANCEL PACKAGE RESONANCES



$$V_{OUT} = V_Y \log \frac{\varepsilon_A}{\varepsilon_B}$$

$$= V_Y \log \frac{V_A g(f)}{V_B g(f)}$$

$$= V_Y \log \frac{V_A}{V_B}$$

Both channels have the same HF resonances and other HF transmission effects g(f), but these are canceled in taking the difference which remains a measure of the RATIO of V_A / V_B

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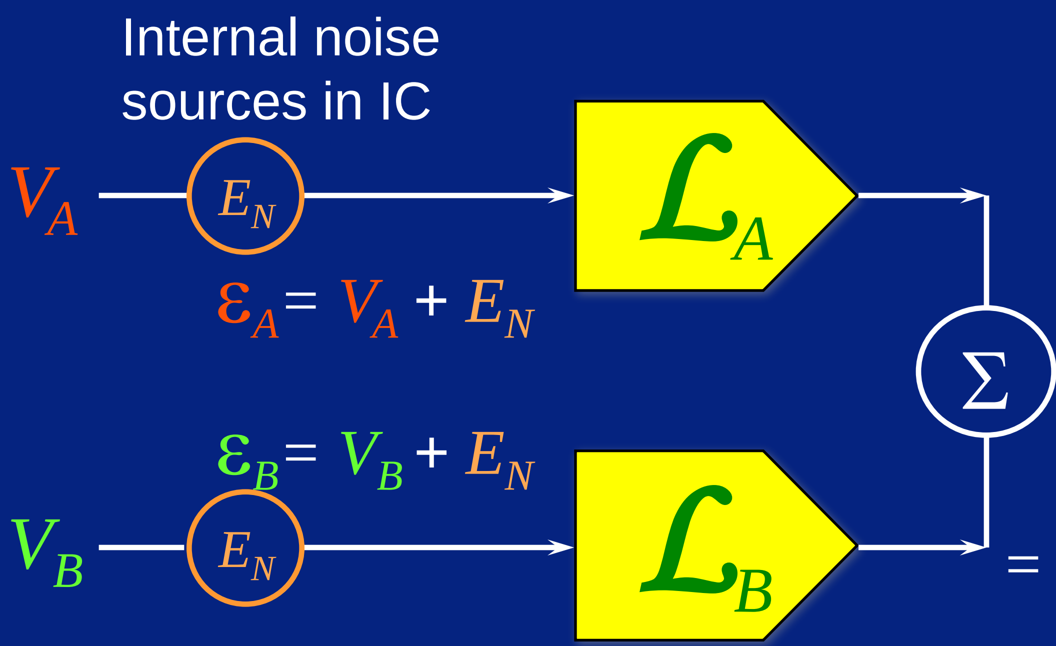
LOW NOISE-INDUCED ERRORS



$$V_{OUT} = V_Y \log \frac{\varepsilon_A}{\varepsilon_B}$$

$$= V_Y \log \frac{V_A + E_N}{V_B + E_N}$$

$$= V_Y \log \frac{V_A}{V_B}\left[1 + E_N \frac{(V_B - V_B)}{V_A V_B}\right]$$

Both channels have the same additive noise at input but its effect is reduced in the output. For example, if $V_A = 3\ E_N$ and $V_B = 2\ E_N$ the measurement error of ratio is only 1.3dB

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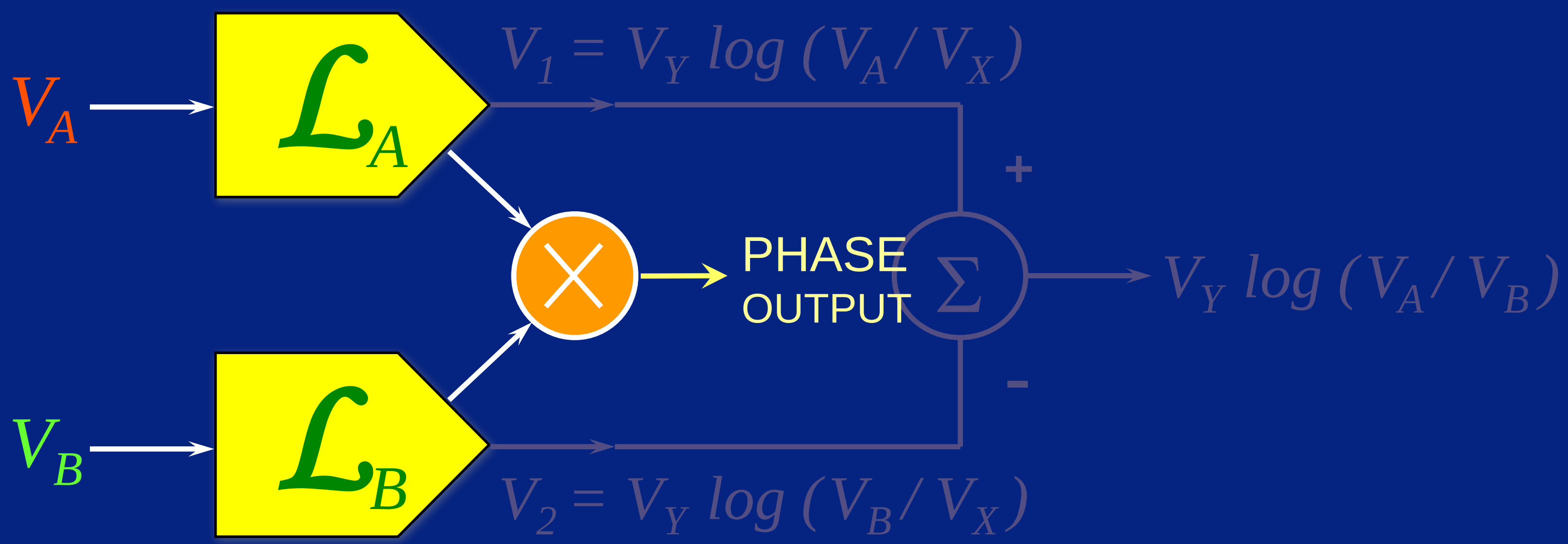
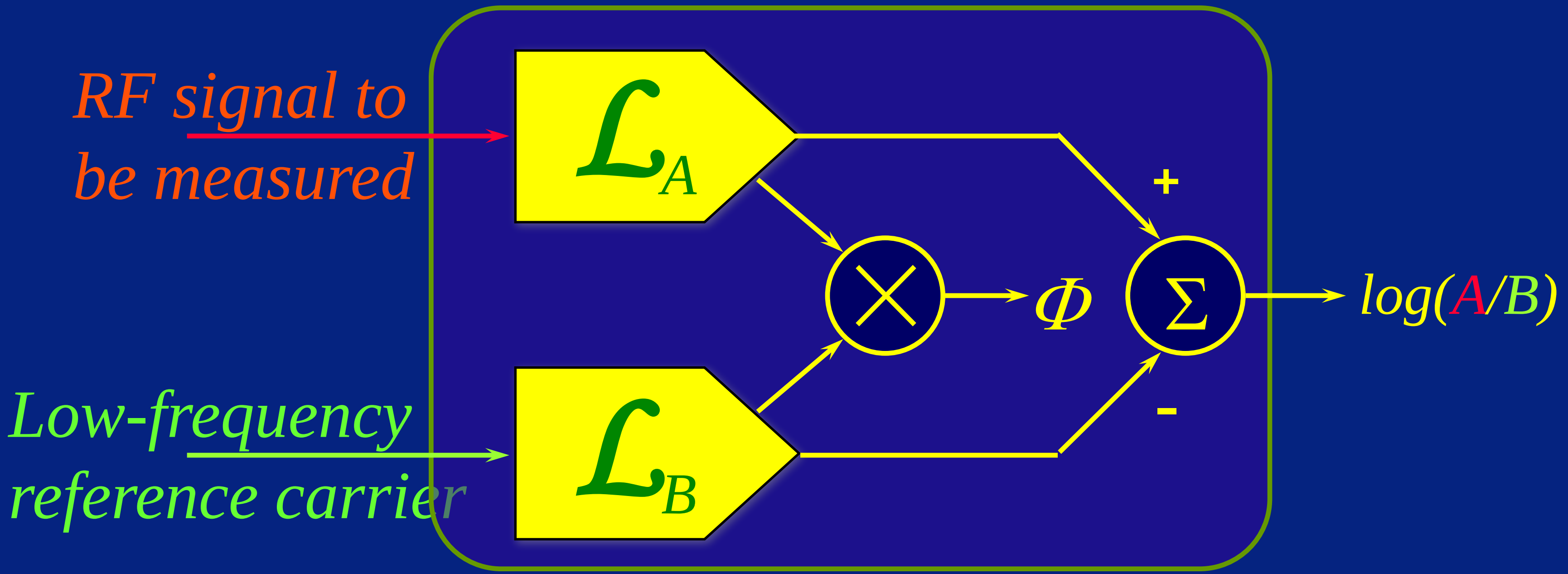
PHASE MEASUREMENT LF-3GHz



Logarithmic amplifiers also provide very high gain and limiting action: using a special type of analog multiplier between the limiter outputs, phase measurements can be made up to 3GHz

APPLICATIONS

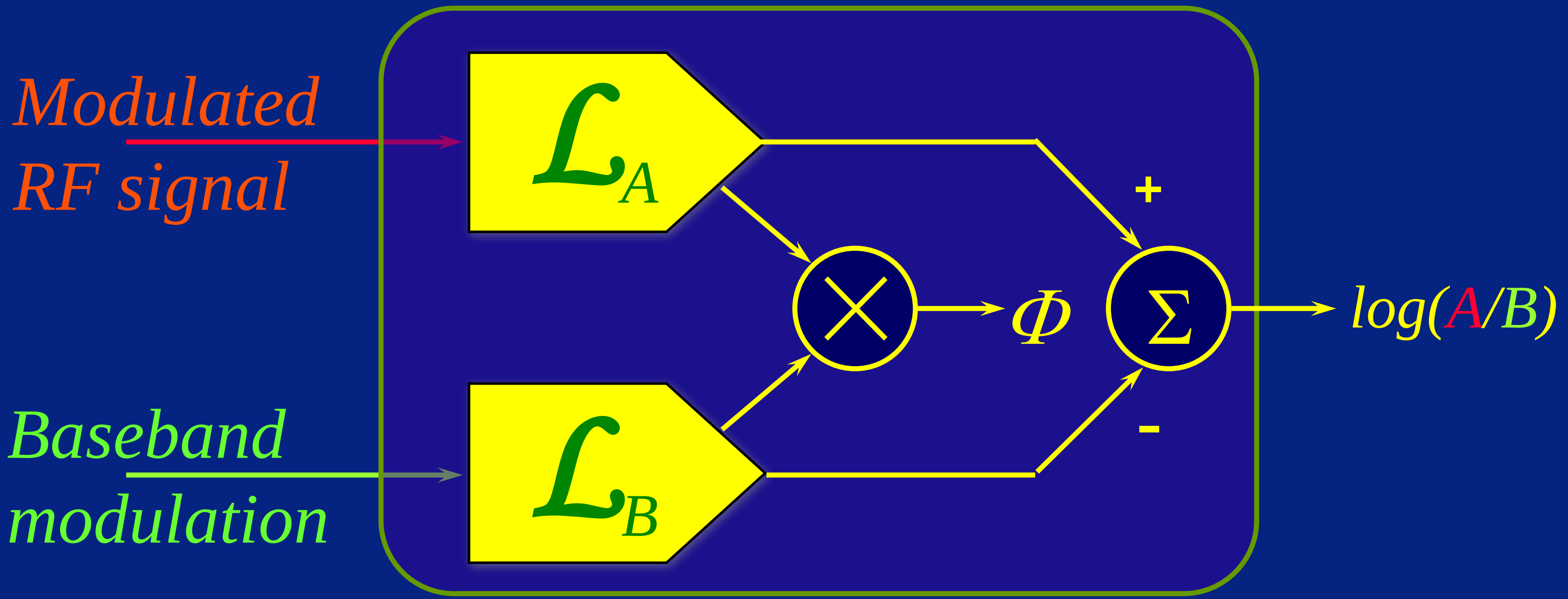
ANALOG DEVICES



In this case, a low-frequency carrier provides a very high calibration reference for the intercept

APPLICATIONS

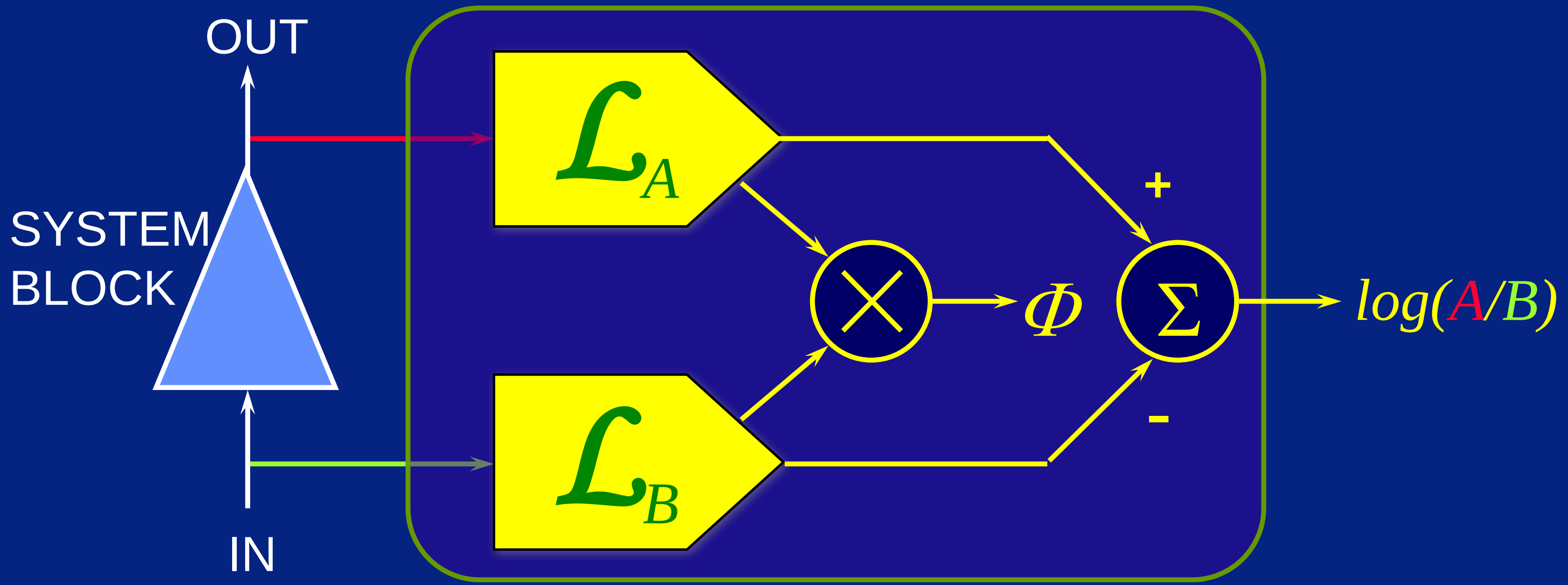
ANALOG DEVICES



Here, the reference is provided by the baseband modulation & system measures conversion gain

APPLICATIONS

ANALOG DEVICES



True gain of system block is measured independent of the actual power levels

ANALOG DEVICES

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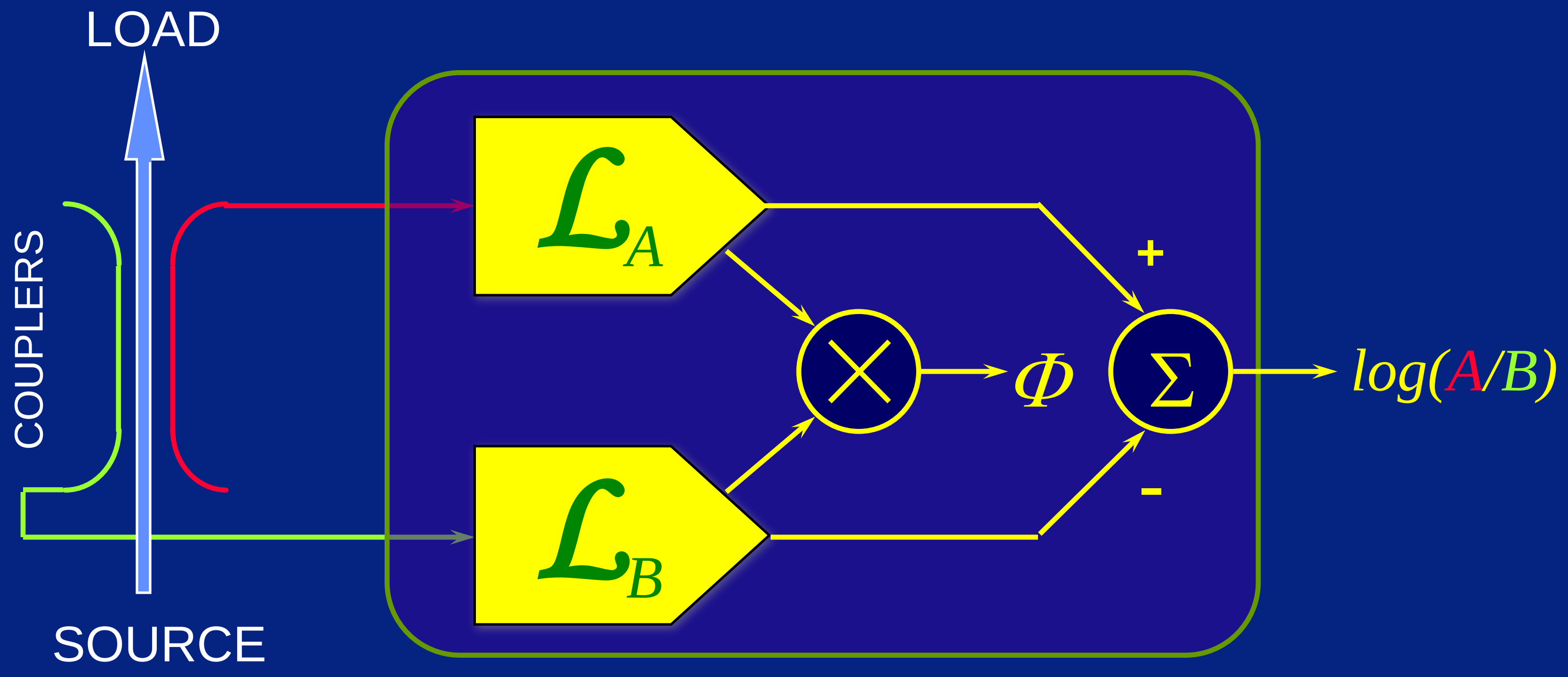
APPLICATIONS



Measurement of return loss independent of power level

ANALOG DEVICES

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SYNCHRONOUS LOG AMP

SYNCHRONOUS LOG AMP

- AN ‘SLA’ IS DUAL LOG-AMPS ACTING IN PARALLEL
- INSTEAD OF ‘SQUARE-LAW’ DETECTORS IT USES ANALOG MULTIPLIERS BETWEEN CORRESPONDING NODES AS SIGNALS PROGRESS DOWN THE CASCADE
- CURRENT-MODE OUTPUTS OF ALL MULTIPLIERS IS SUMMED, AND THIS VARIABLE IS CONVERTED BACK TO THE VOLTAGE DOMAIN
- NUMEROUS APPLICATIONS: LOWER EFFECTIVE INPUT NOISE; TUNABLE TO SINGLE FREQUENCY; $\sinh^{-1}$; etc

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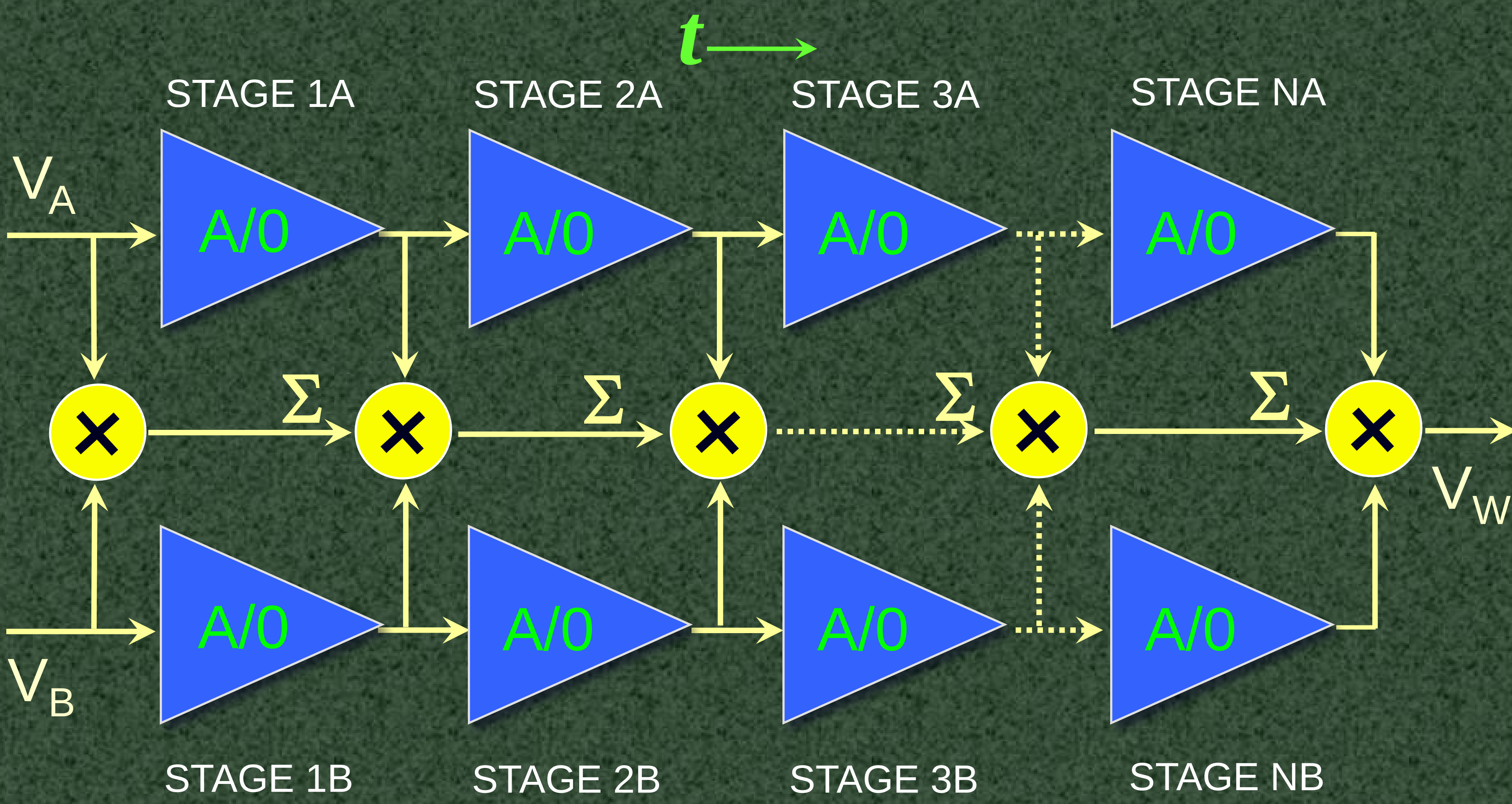
SYNCHRONOUS LOG AMP



2 × N STAGES, N IS TYPICALLY 12

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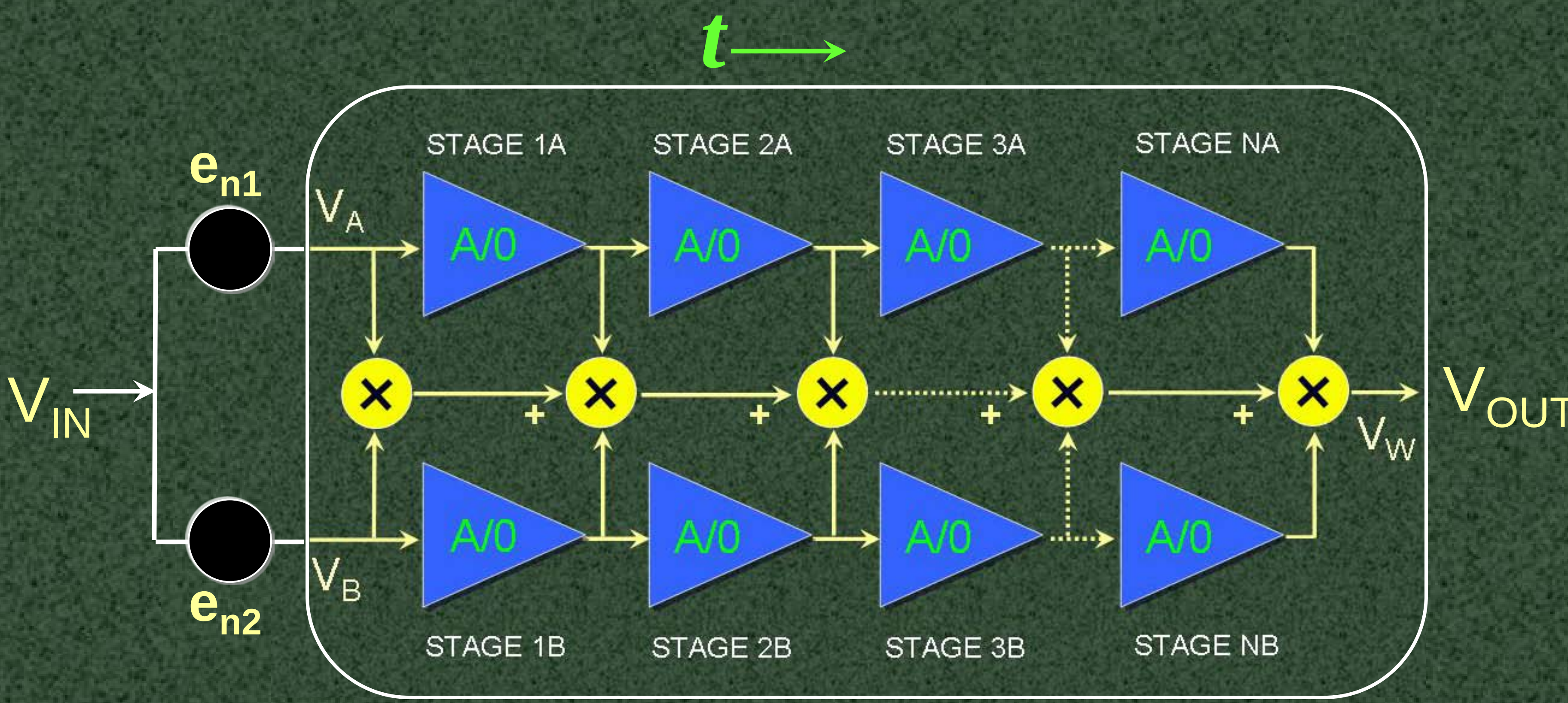
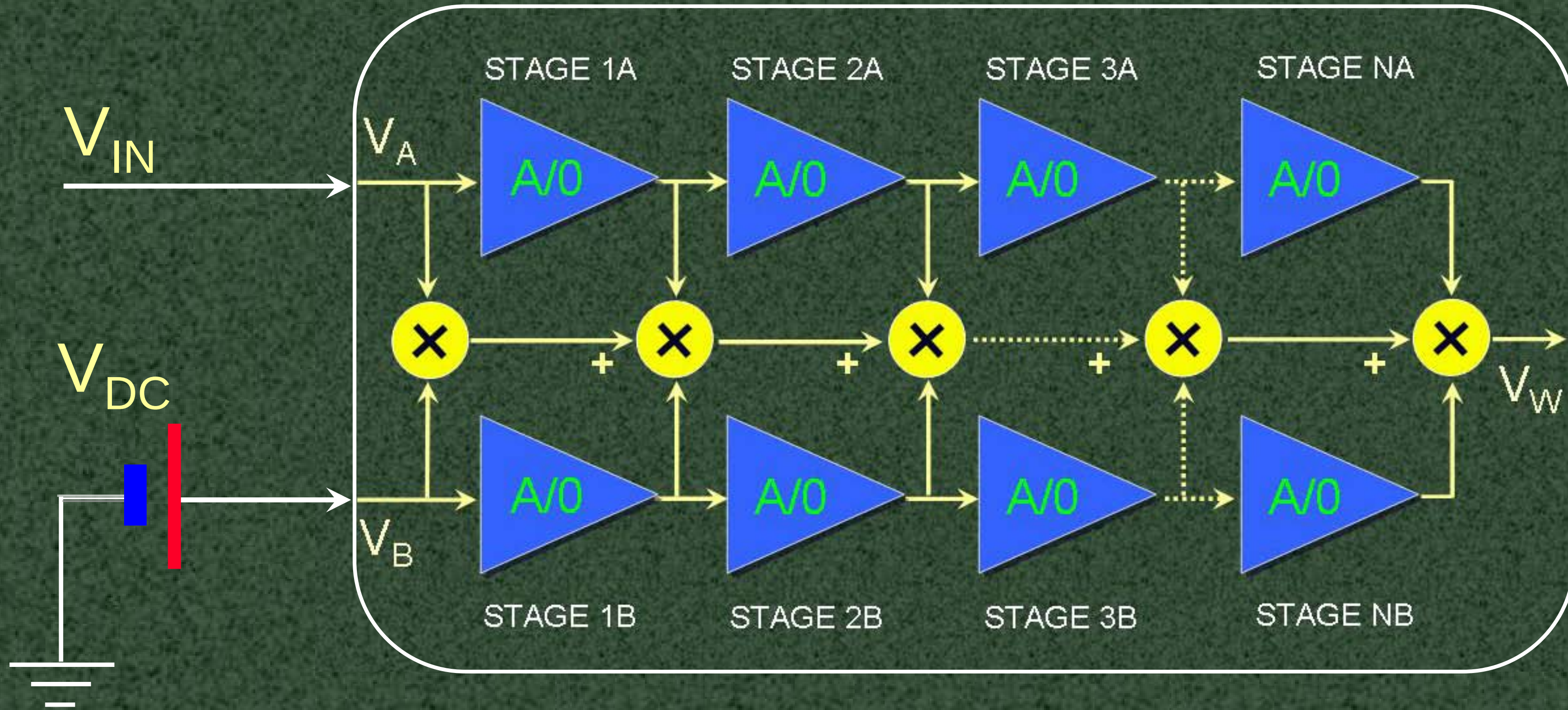
FIRST APPLICATION

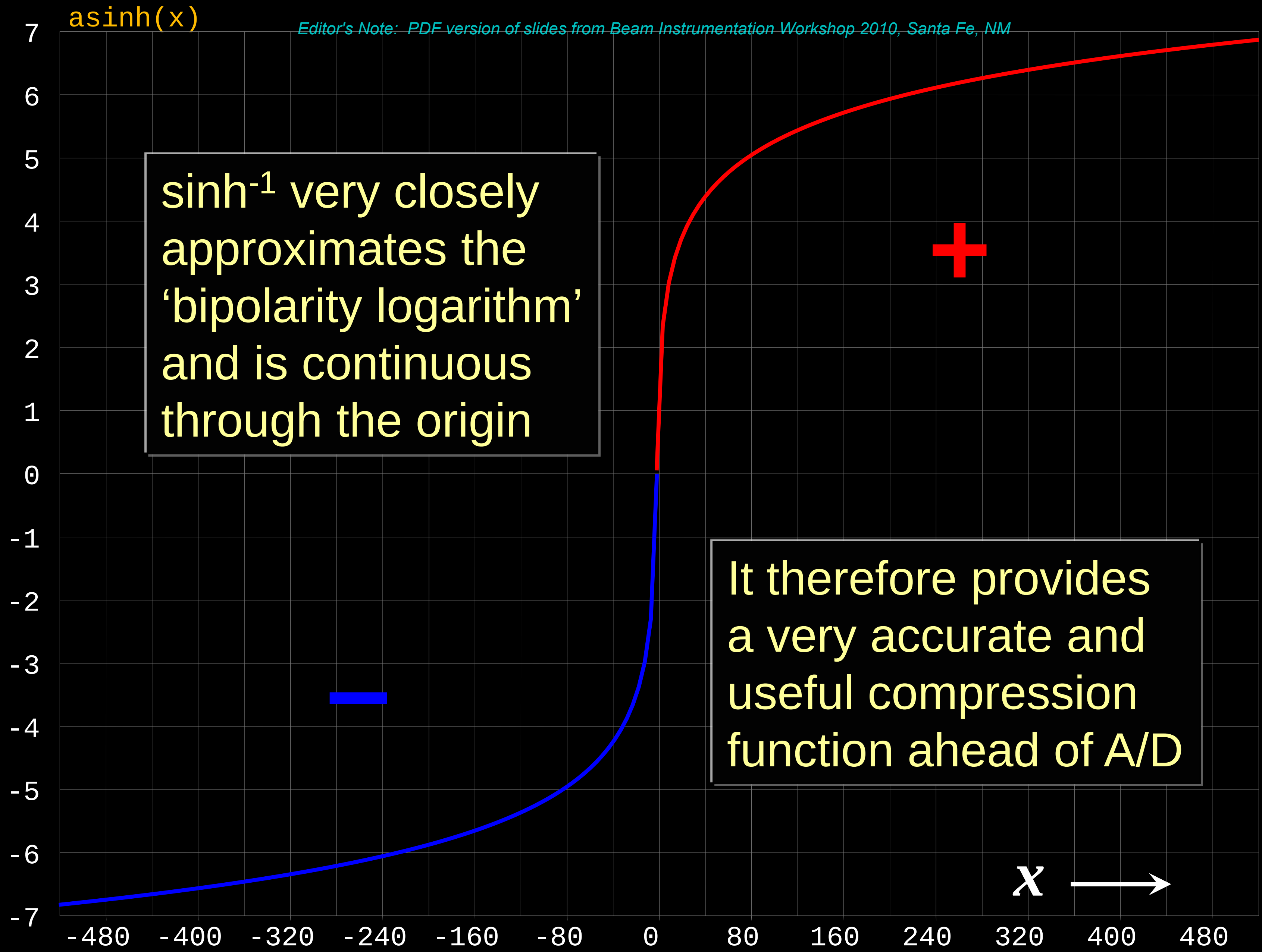


NOISE SOURCES e_{N1} and e_{N2} ARE UNCORRELATED SO THEIR CROSS-PRODUCT AVERAGES TO ZERO

SECOND APPLICATION



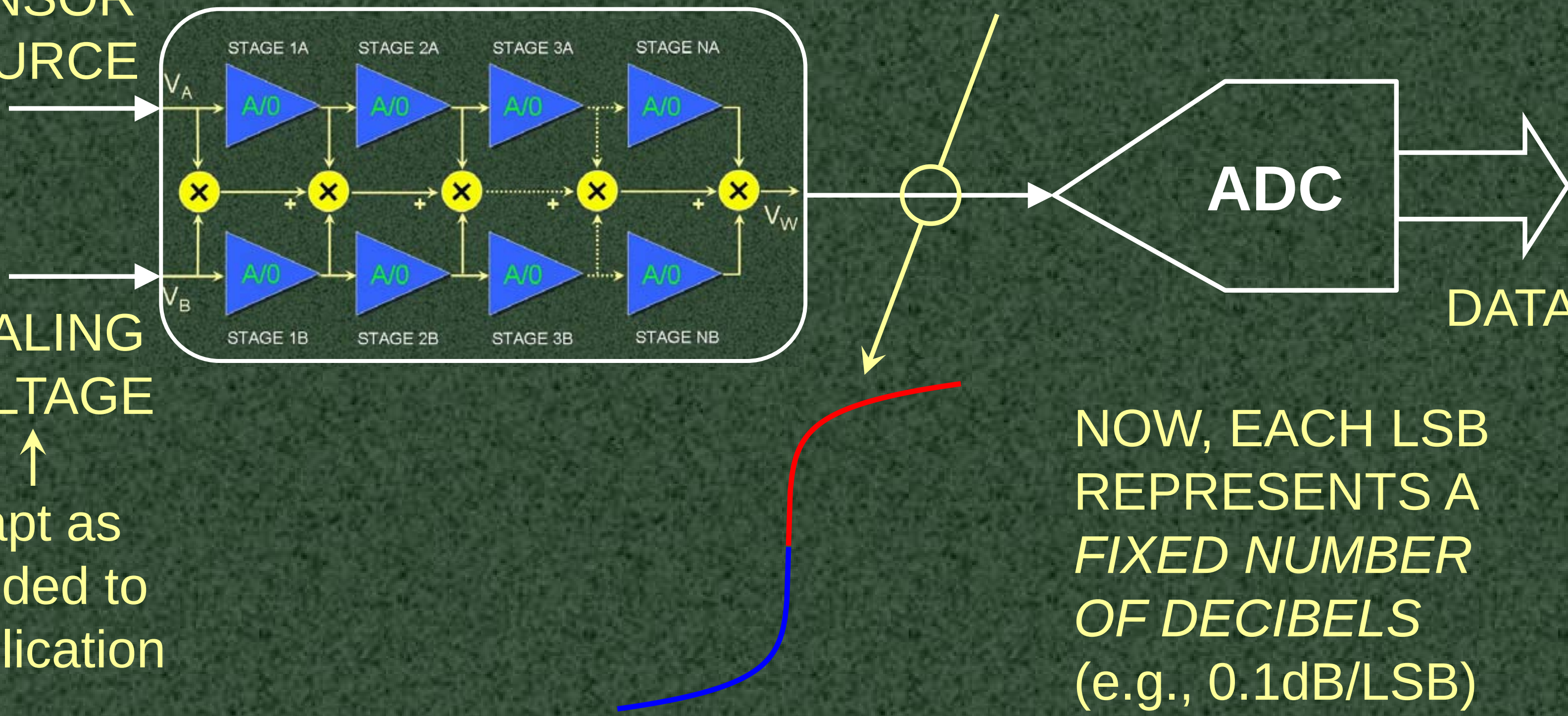
THE FIXED VOLTAGE V_{DC} TURNS THIS LOG AMP INTO A NON-DEMODULATI NG "$\sinh^{-1}$" MACHINE


asinh(x)
7
6
5
4
3
2
1
0
-1
-2
-3
-4
-5
-6
-7
-480 -400 -320 -240 -160 -80 0 80 160 240 320 400 480
+
–
x →
sinh⁻¹ very closely approximates the ‘bipolarity logarithm’ and is continuous through the origin
It therefore provides a very accurate and useful compression function ahead of A/D

PRE-A/D COMPRESSION

SIGNAL/ SENSOR SOURCE

LOGARITHMICALLY COMPRESSED AND SCALED SIGNAL



SCALING VOLTAGE

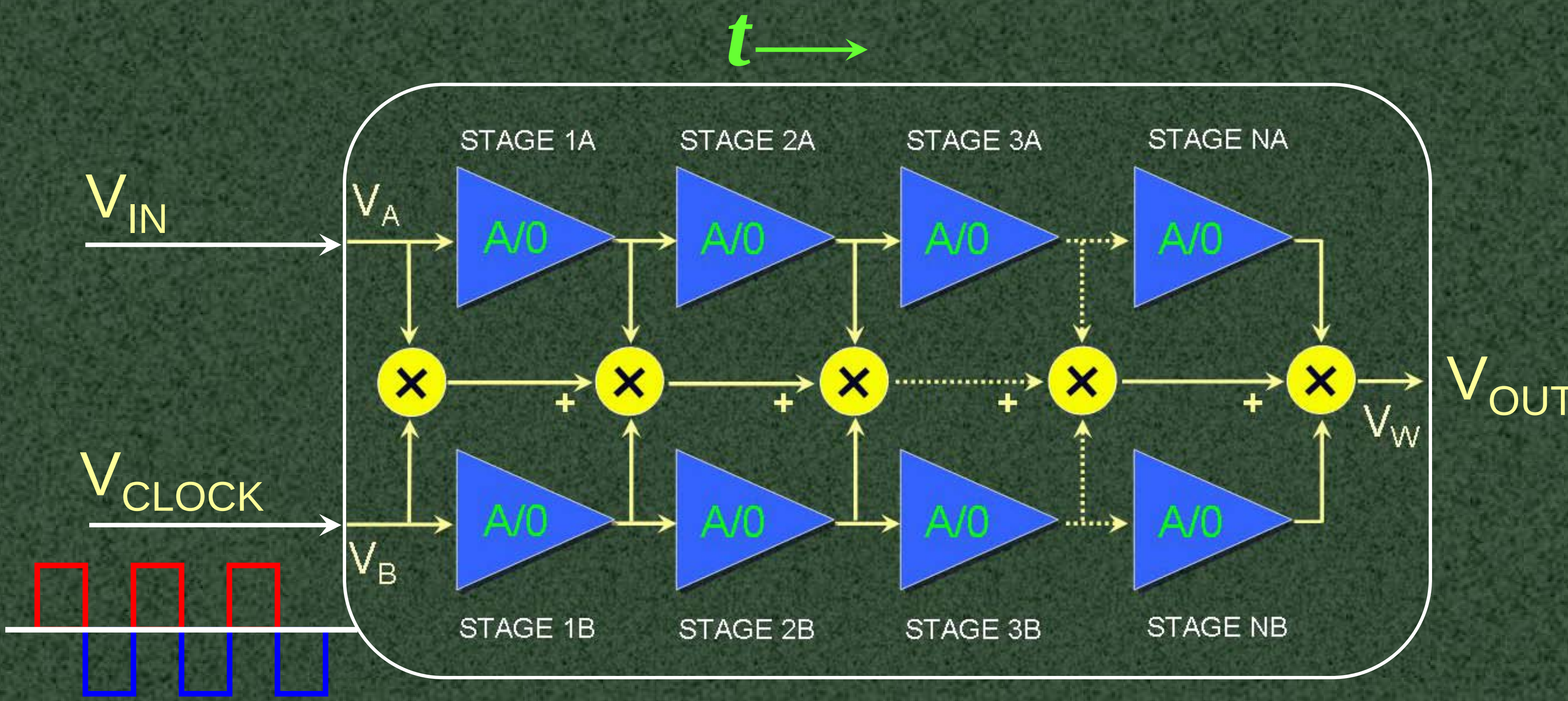
↑

Adapt as needed to application

NOW, EACH LSB REPRESENTS A *FIXED NUMBER OF DECIBELS* (e.g., 0.1dB/LSB)

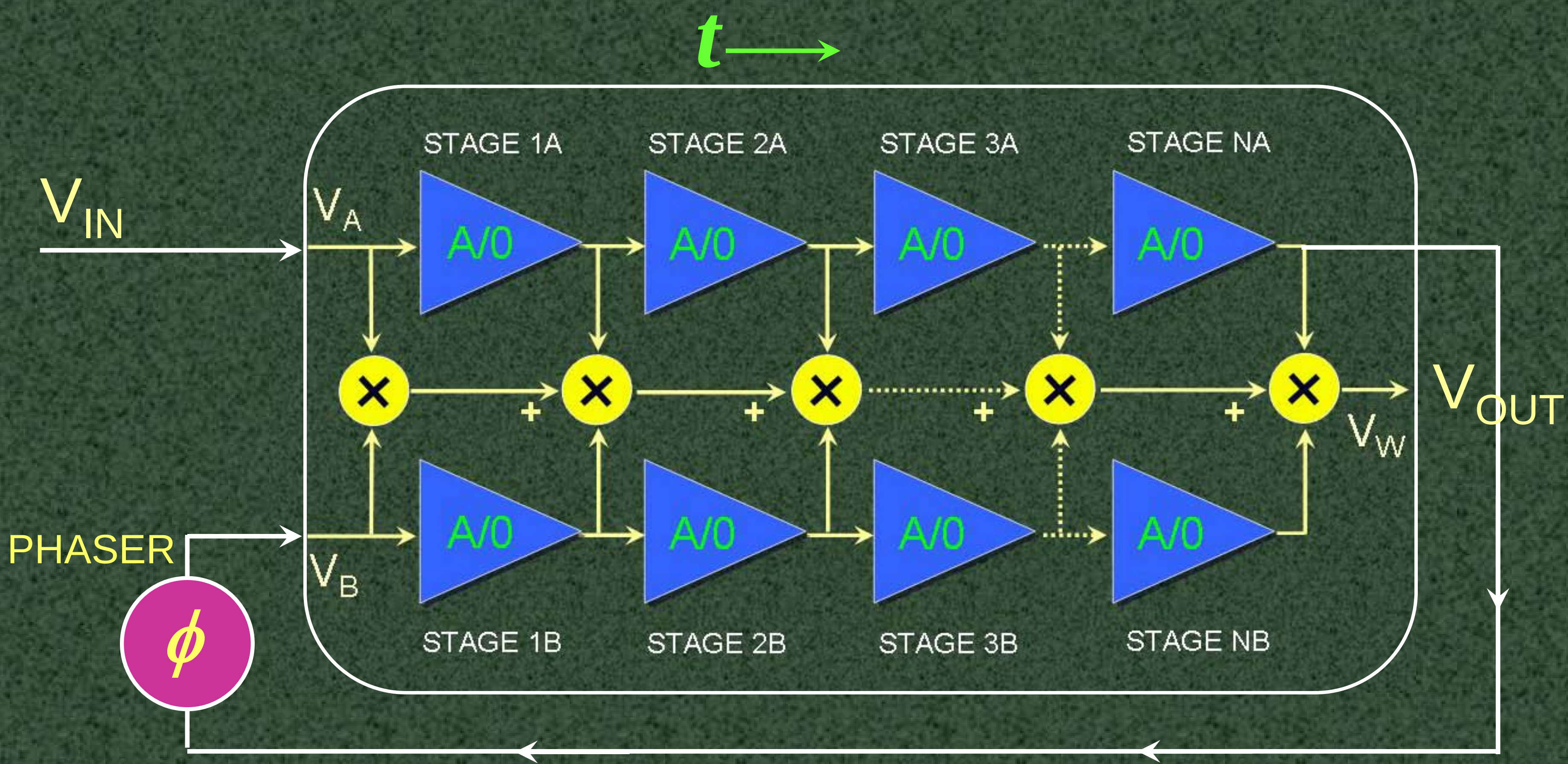
ANALOG DEVICES

THIRD APPLICATION



THE CLOCK VOLTAGE TURNS THIS LOG AMP INTO A SYNCHRONOUS DEMODULATOR

FOURTH APPLICATION



THE CLOCK VOLTAGE IS RECOVERED FROM THE LAST LIMITER OUTPUT: SYSTEM BECOMES A SELF-CLOCKED ‘HOMODYNE’

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DLVA

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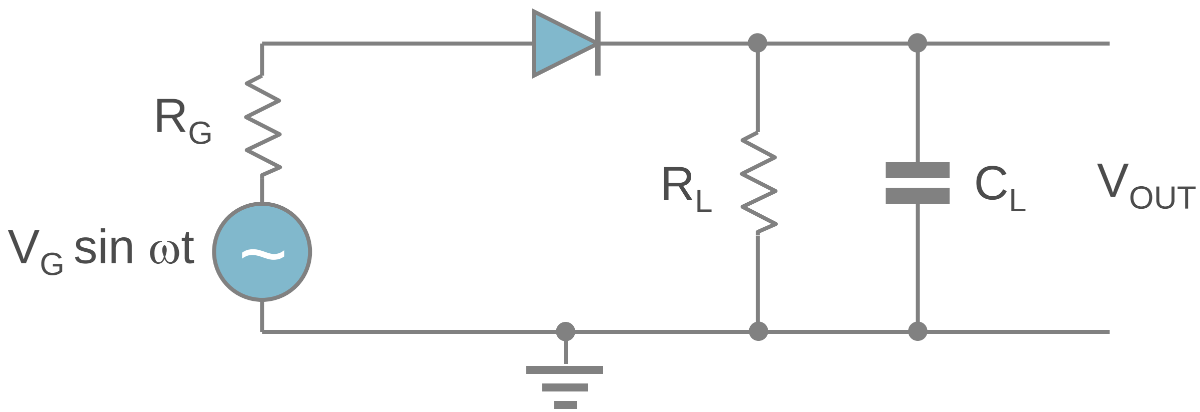
A PRIMITIVE SCHOTTKY DETECTOR

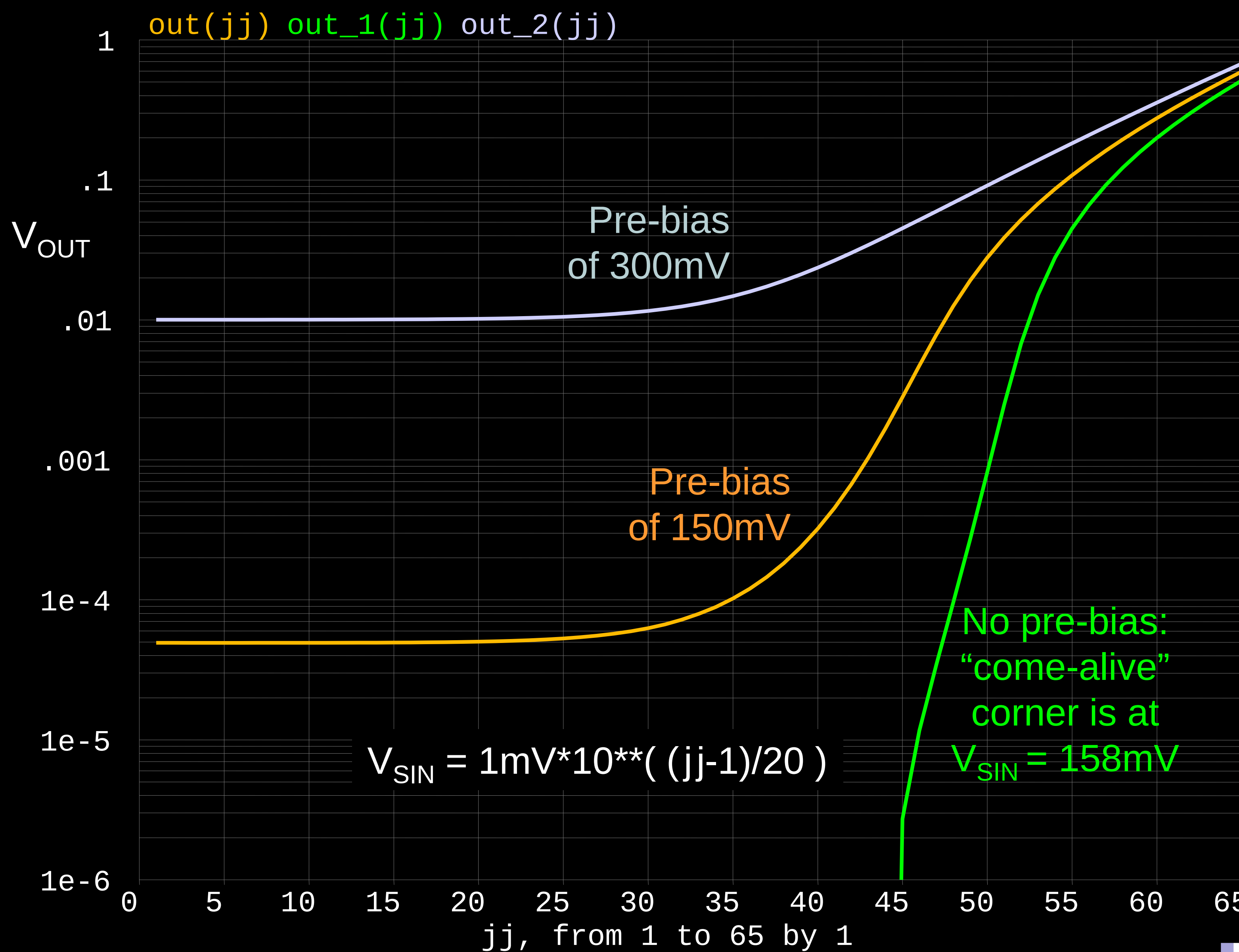


The great appeal of this Schottky-diode detector is its inherently wide bandwidth, passing effortlessly through the SHF band (3-30GHz) and still working very well into the EHF band (30-300GHz).

However, it suffers from extremely poor sensitivity, severe nonlinearity, and temperature dependence.

Thus, the challenge is to find ways to persuade some form diode detector to be free of these limitations, with the aim of providing dynamic range of at least 50dB.

AMPLITUDE RESPONSE, fsin = 1GHz, 1mV to 1.58V in 1-dB steps



ANALOG DEVICES

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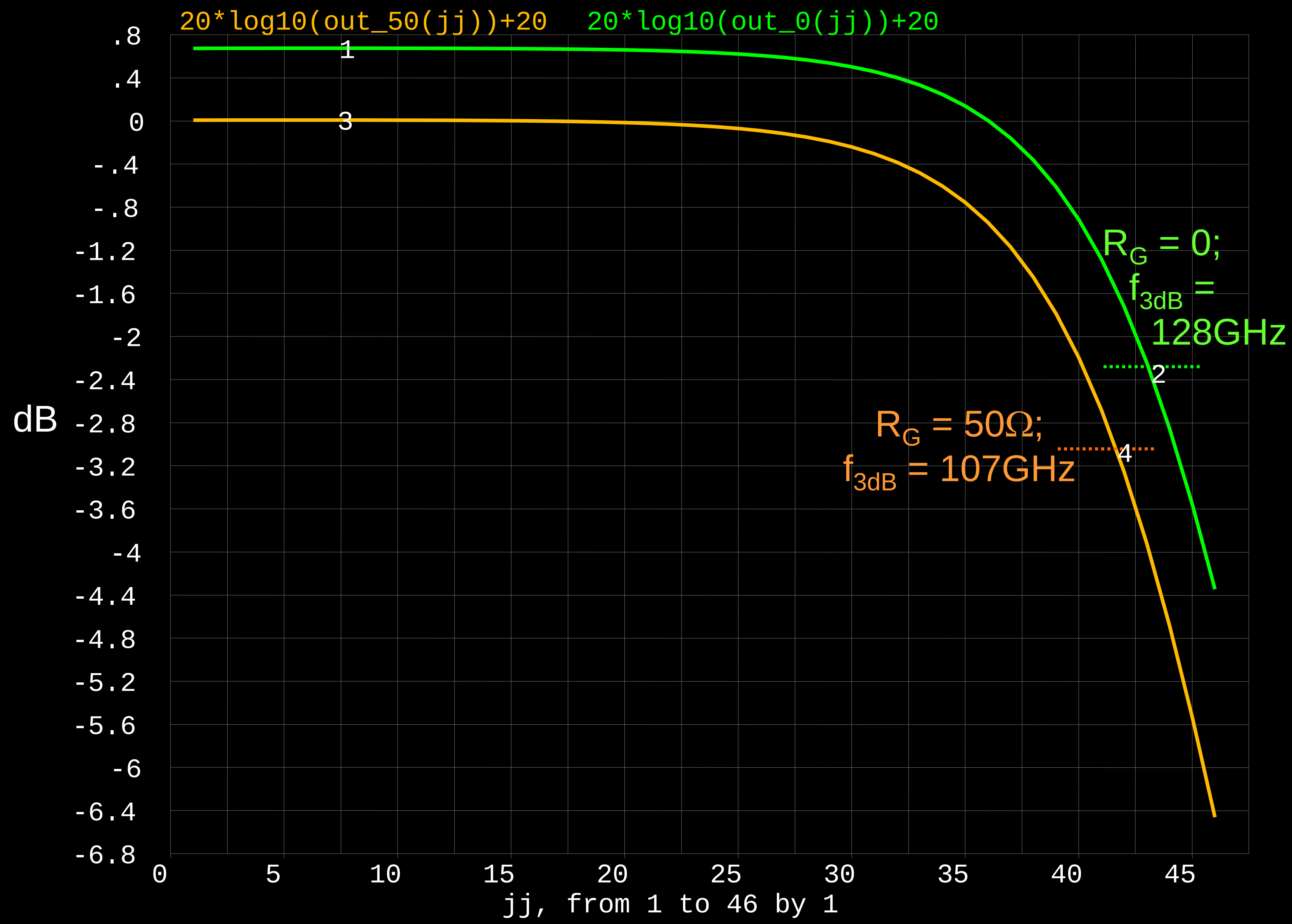
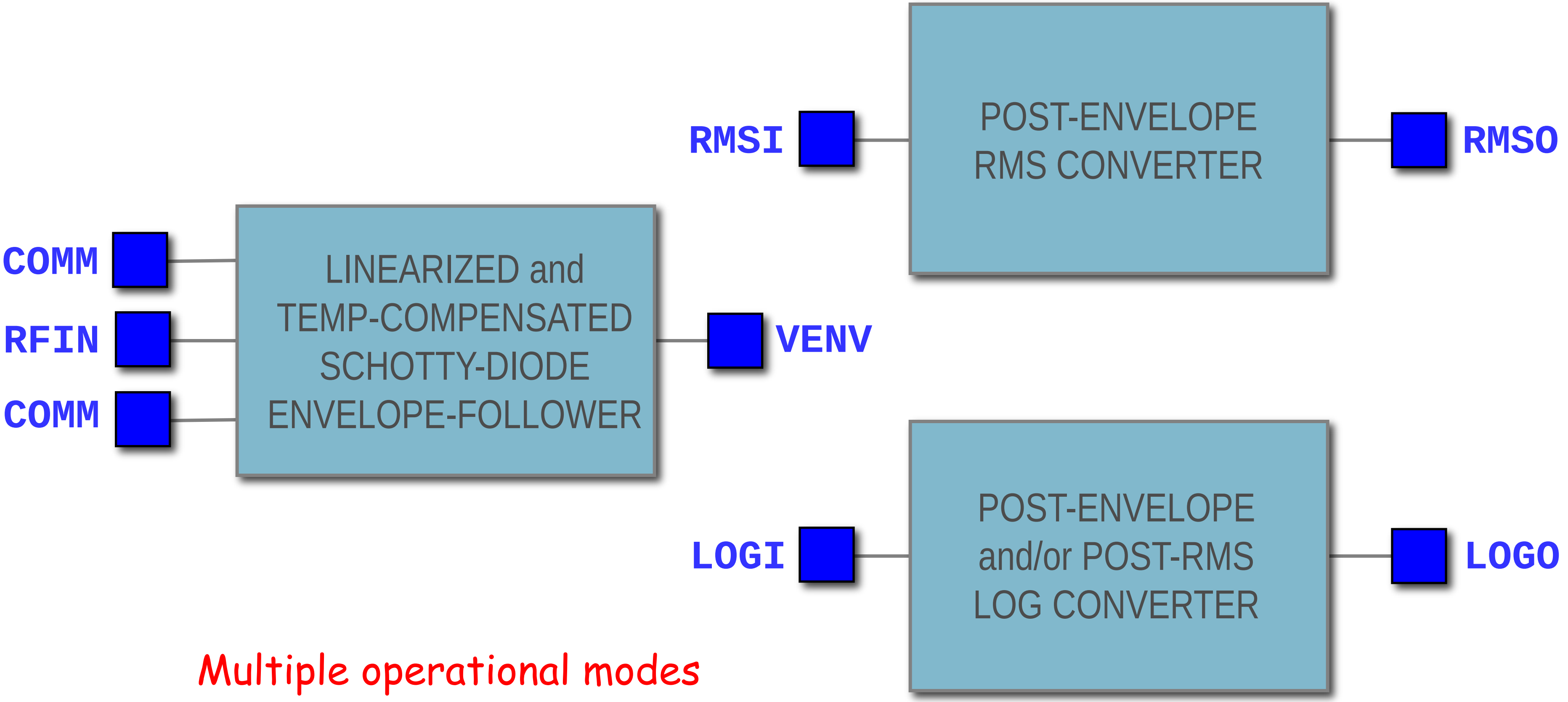
FREQUENCY RESPONSE, Vsin = 300mV, 1GHz to 316GHz



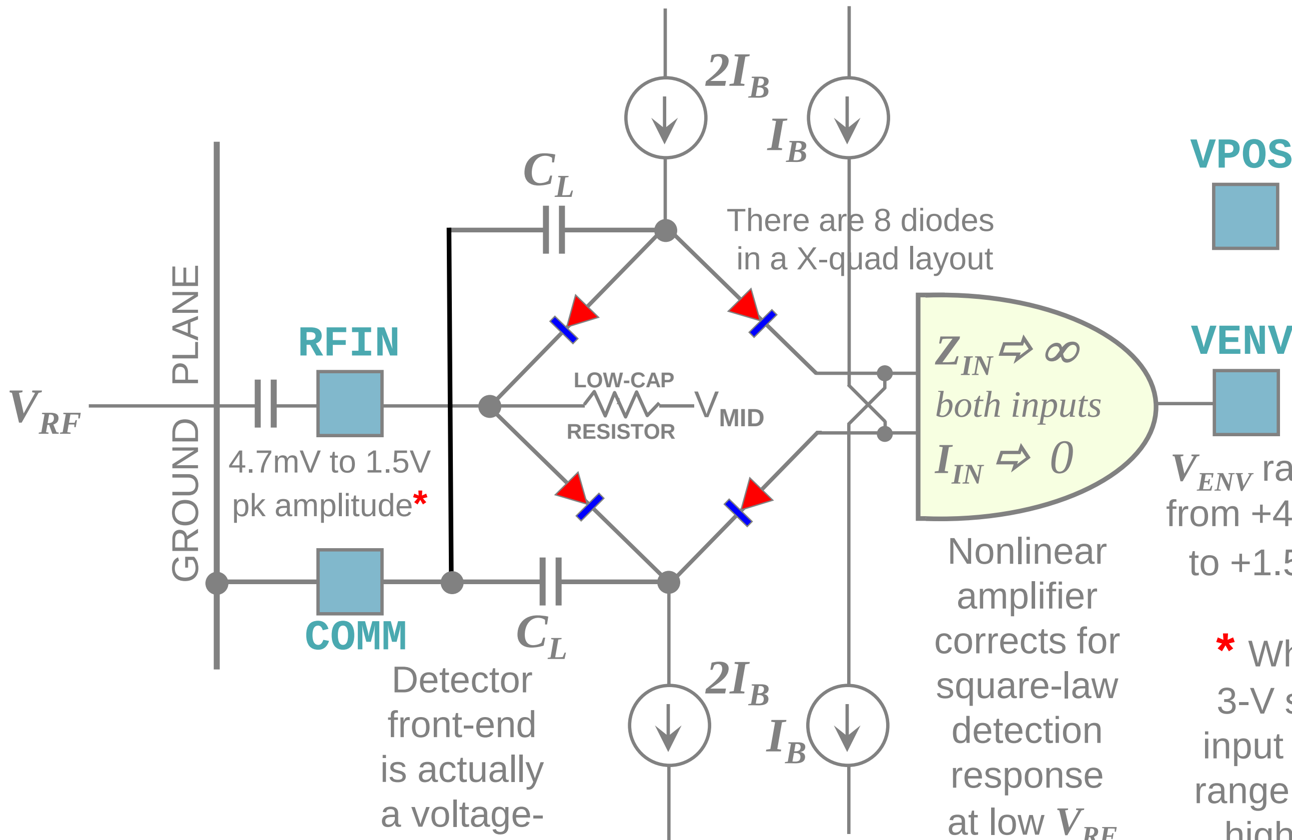
DLVA: BASIC BLOCK DIAGRAM



Multiple operational modes

ANALOG DEVICES

THE FRONT-END



Detector front-end is actually a voltage-doubling circuit

Nonlinear amplifier corrects for square-law detection response at low V_{RF}

V_{ENV} ranges from +4.7mV to +1.5V *

* When using a 3-V supply; the input and output range can be 4dB higher using a 5-V supply

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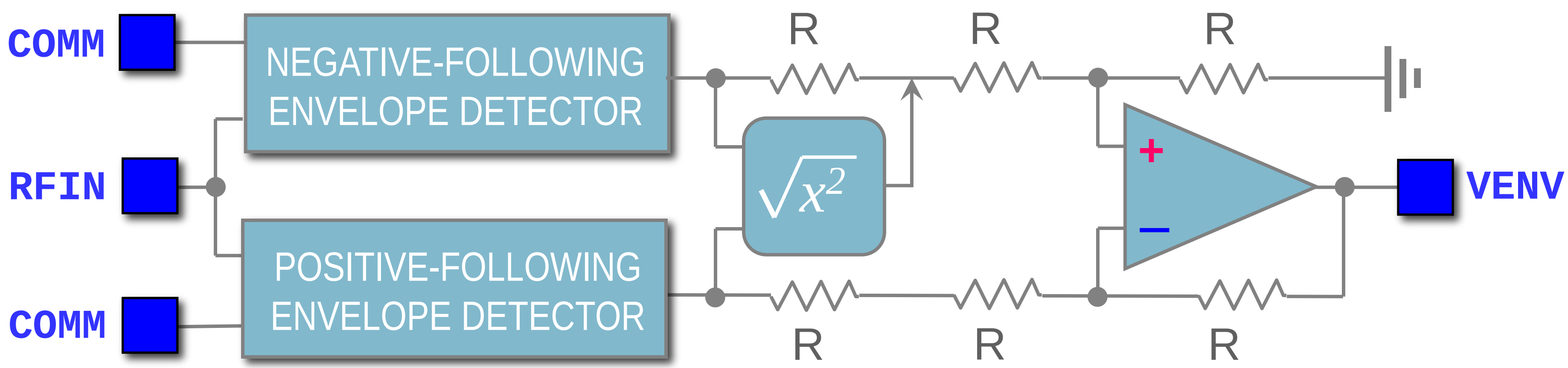
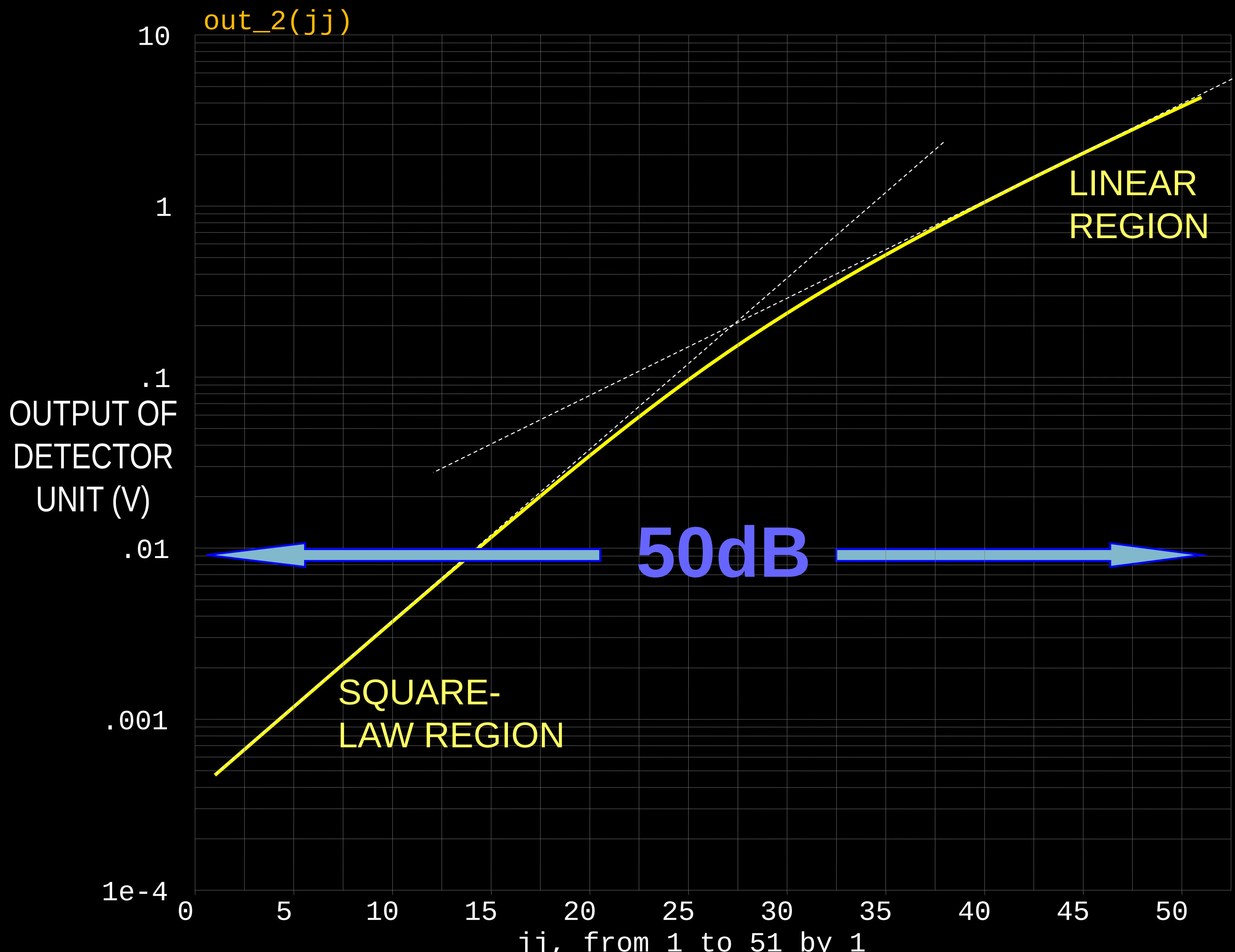
FRONT-END PROCESSING



Because the input detectors provide voltage-doubling, the gain of the output section is 0.5.

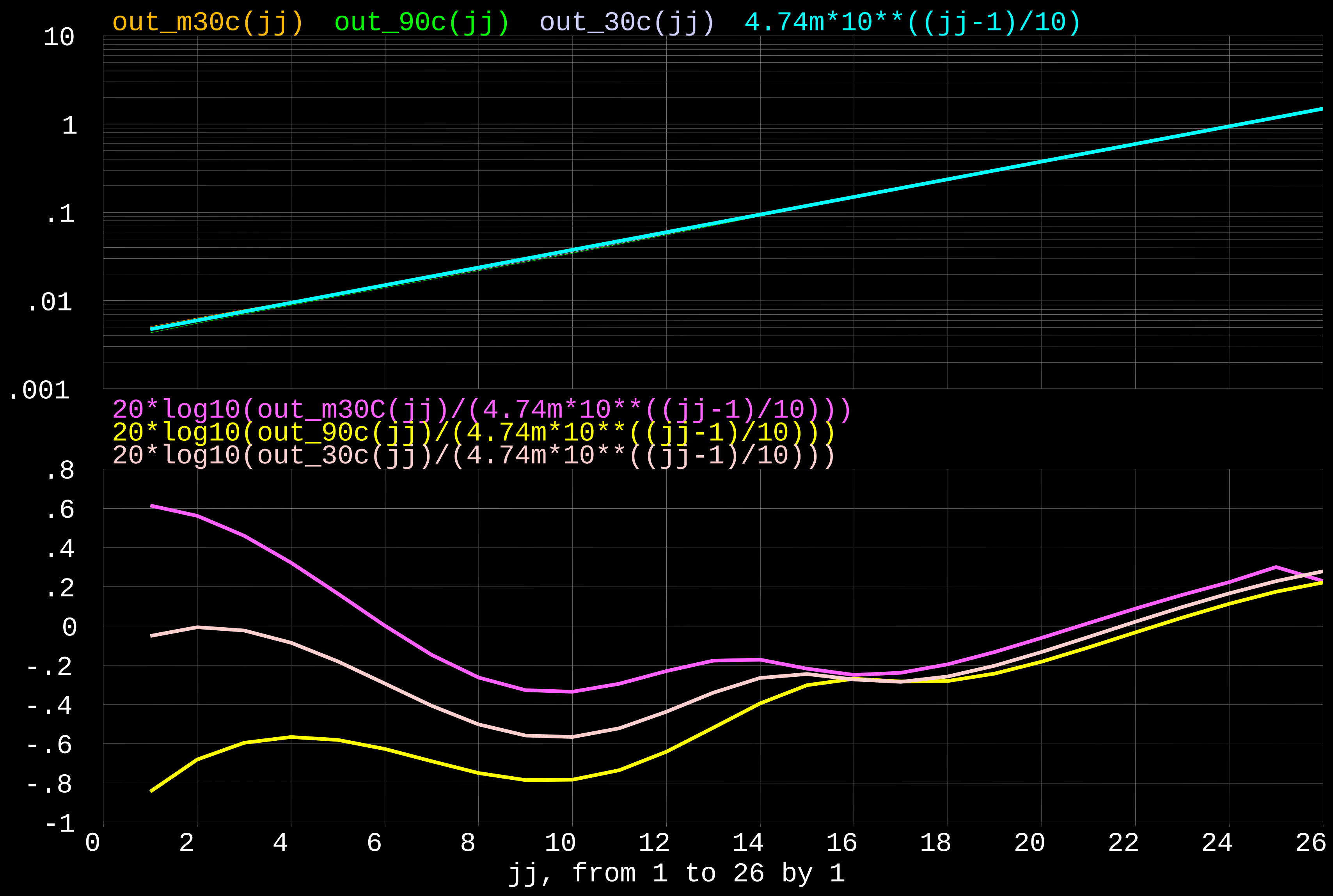
The square-rooting cell is a particular sort of translinear circuit, having the double-amplitiude differential voltage applied to its ultra-low-offset transconductance stage and the resulting current is used to generate the needed correction current to the output amplifier. To maintain fast response time, this current is kept fairly high and then attenuated in the resistor string.

Editor's Note: PDF version of slides from Beam Instrumentation Workshop 2010, Santa Fe, NM



ANALOG
DEVICES

An overnight run of March 20-21, 2010, checking temperature error at fsin = 500MHz, after small adjustment to SQRT scaling. Using inputs from 4.74mV to 1.500V (=50.006dB).



ANALOG DEVICES

Editor's Note: PDF version of slides from Beam Instrumentation Workshop 2010, Santa Fe, NM

tanh ***sech***

TRUST THE MATHEMATICAL TRANSISTOR

exp ***log***

cosh ***sinh***