• Notes Set 11: Noise of miscellaneous devices

- Schottky and PN junction diodes. transmission lines. antennas
- antenna noise temperature. Temperature of the sky. Friss formula in temperature form.

ECE ___ Notes set 11

We have now developed noise models of:

* bipolar transistors

* Field-effect transistors.

* Resistors

We must also ~~do~~ consider noise models of

* diodes

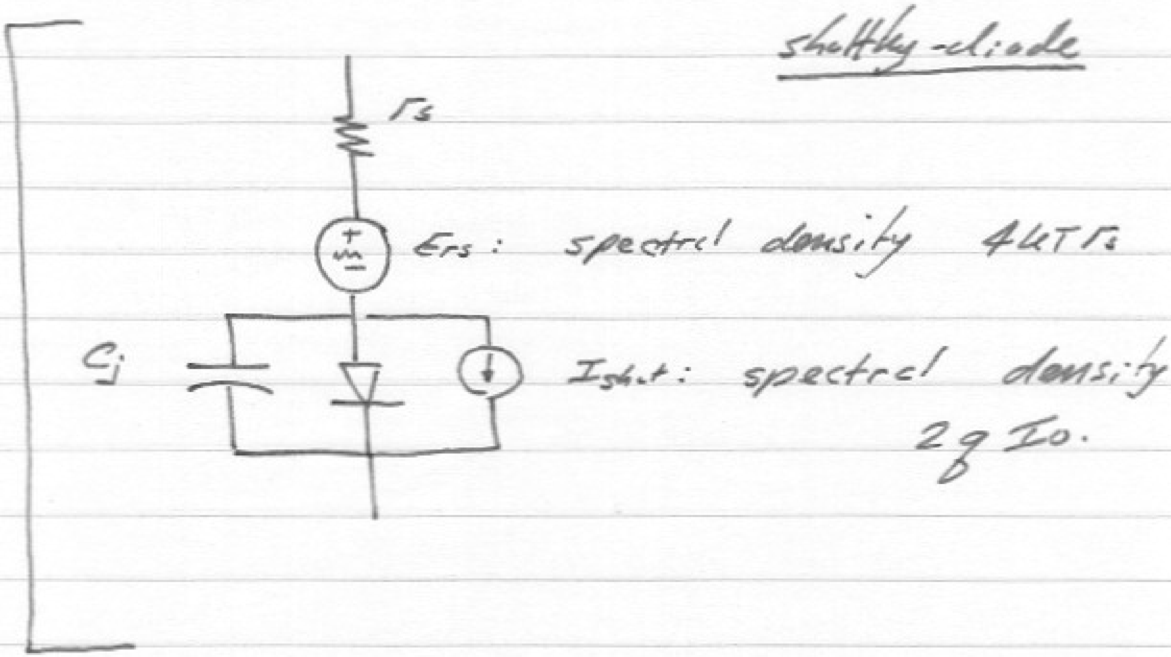
* Antennas

* Transmission Media:

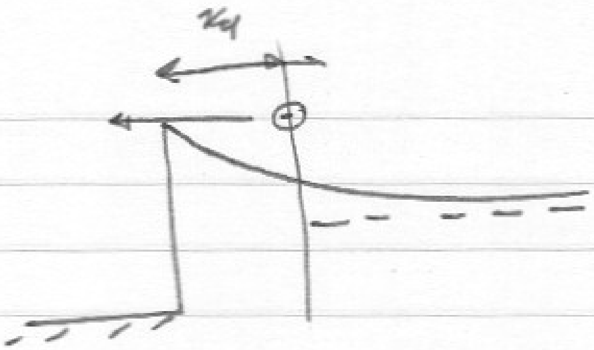
Diodes:

We have largely covered these, but a few corrections remain: The noise current has a spectral density $2q(I_{dc} + 2I_o)$, but

shottky-diode



real diodes have series resistance & junction capacitance which must be added to the model.

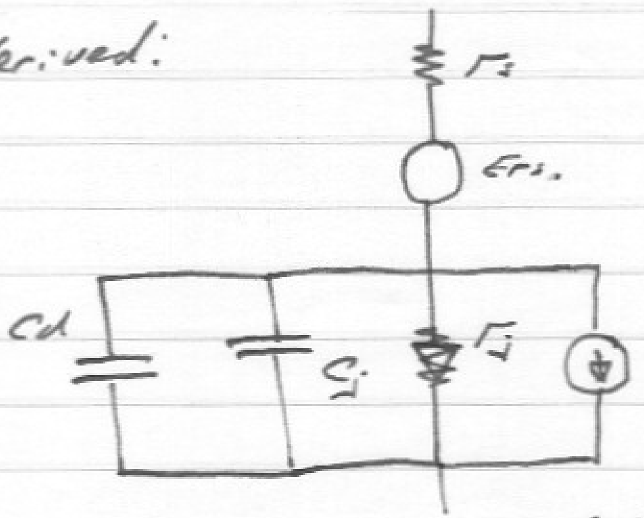


schottky diodes have a depletion region of width x_d, giving a transit time in forward bias of $\tau_T \cong K\, x_d / v_{thermal}$*, which will change the shot noise spectral density to

$$2qI_0 \cdot \left[\frac{\sin \omega\tau_c}{\omega\tau_c}\right]^2 \cong \frac{2qI_0}{1+\omega^2\tau_c^2}$$

* this can become a long discussion! Avoid it.

For P-N. junction diodes in the short-base limit, and for thin depletion regions, the diode model includes a diffusion capacitance as derived:



I_{shot}: $2q\,(I_{dc} + 2I_0)$

with $C_d \cdot r_j = W_{base}^2 / 2D$

* For long-base diodes r_j & C_d become frequency-dependent, as does I_{shot}. This was given earlier.

* For long depletion regions (P-I-N) diodes, the device analysis becomes lengthy whether for noise or just small-signal. See Sze.

What about Inductors, Capacitors, and transmission lines?

Ideal , , and transmission lines have ~~mostly~~ no means of dissipating power. Their terminal impedance is pure imaginary, so the available noise power is zero.

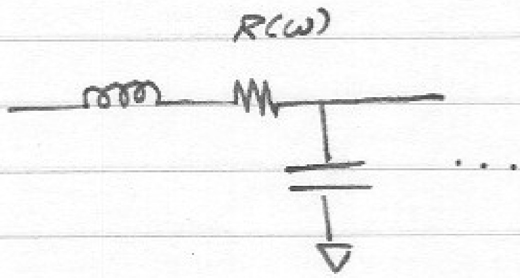
Real ~~I~~ Inductors or Capacitors have loss. Regardless of the physical origin of this loss, it can be modelled by:

... where the series or shunt resistance may be frequency-dependent. In all cases that resistance has thermal noise with available power kT, where T is the local temper-ature.

Transmission-Lines (wires)

require a bit more attention. 3 causes of loss

1) Conductor resistance



$R(\omega)$ is approximately of the form $R_{dc} + K\sqrt{f}$ where the latter term is due to the skin effect. $R(\omega)$ can be simply treated as any other resistor, & has available noise power KT.

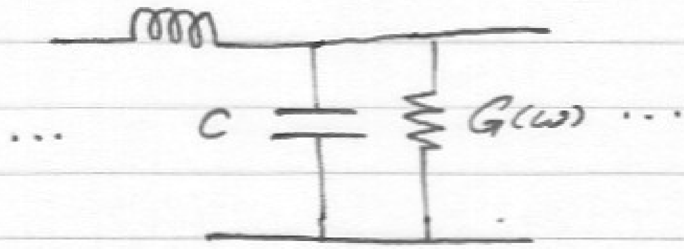
2) Dielectric losses:

$$\epsilon = \epsilon_r \epsilon_0 + j\, \epsilon_{ir} \epsilon_0$$

$\epsilon_{ir} / \epsilon_0$ is called the loss-tangent.

...This is power lost in polarizing & de-polarizing the dielectric..

usually modelled as such:



where $G / j\omega C = \epsilon_r / j \epsilon_{ir}$

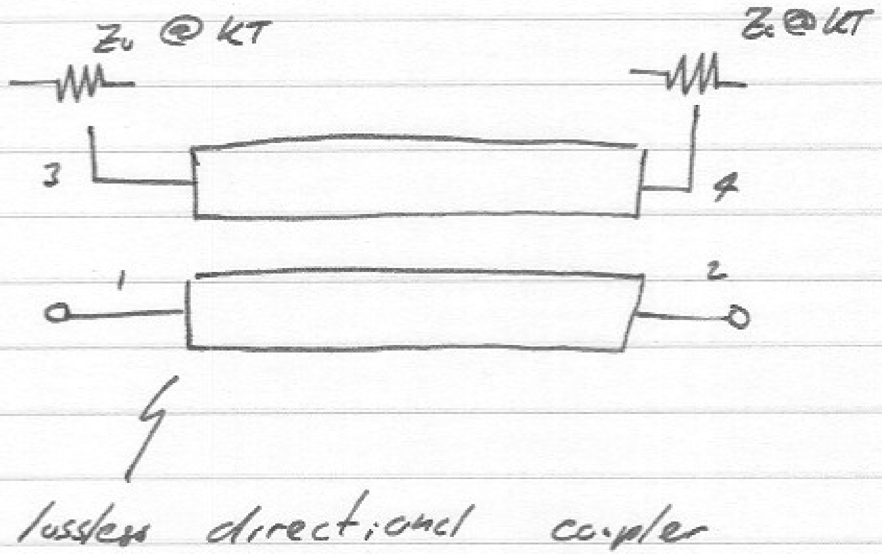
and G has available power kT.

Fortunately $G(\omega)$ is usually negligible compared to skin loss.

3) <u>open</u> transmission lines have radiation losses if <u>completely</u> open, or "mode coupling losses" if the open line is contained within a larger closed package. This is radiation, and the line acts as a transmitting antenna. <u>By</u> reciprocity the line also acts as a receiving antenna, and generates thermal noise power from the thermal radiation of the local environment.

~~A wave approach handles this most easily.~~

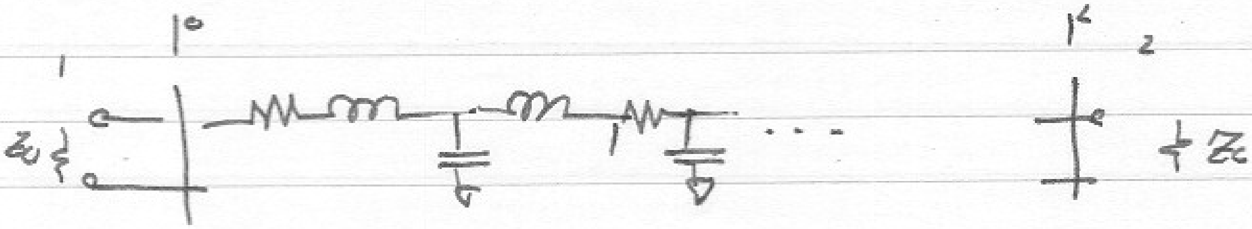
It is tempting to model the transmission line for noise thus:



such that all ports have impedance Z_0 & the directional coupler is picked so that the port 1-2 coupling is equal to the line loss.

But the sources at ports 3 & 4 are not generally uncorrellated

This can be seen for skin loss:



If we call E_{n1} and E_{n2} the voltage wave amplitudes coupling to ports 1 & 2

$$E_{n1} := \int_0^L E_n(z)\, e^{-j\beta z}\, e^{-\alpha z}\, dz$$

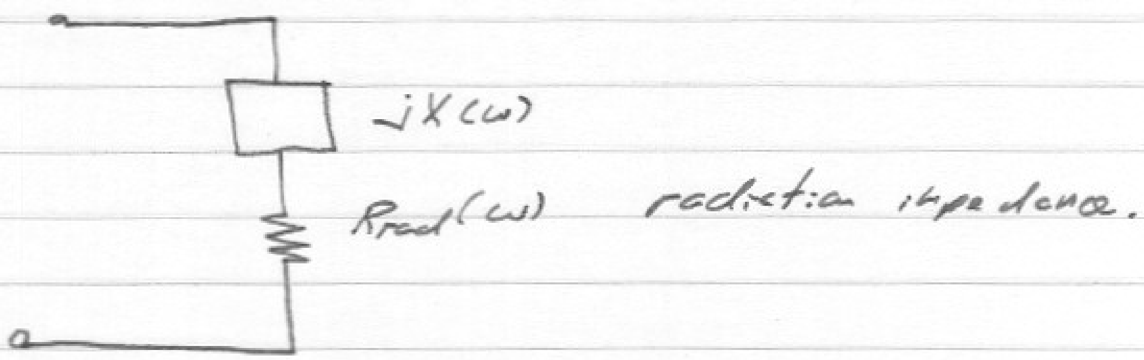
$$E_{n2} := \int_0^L -E_n(z)\, e^{-j\beta(L-z)}\, e^{-\alpha(L-z)}\, dz$$

where $E_n(z)$ has spectral density $4kT\,R(z)$.

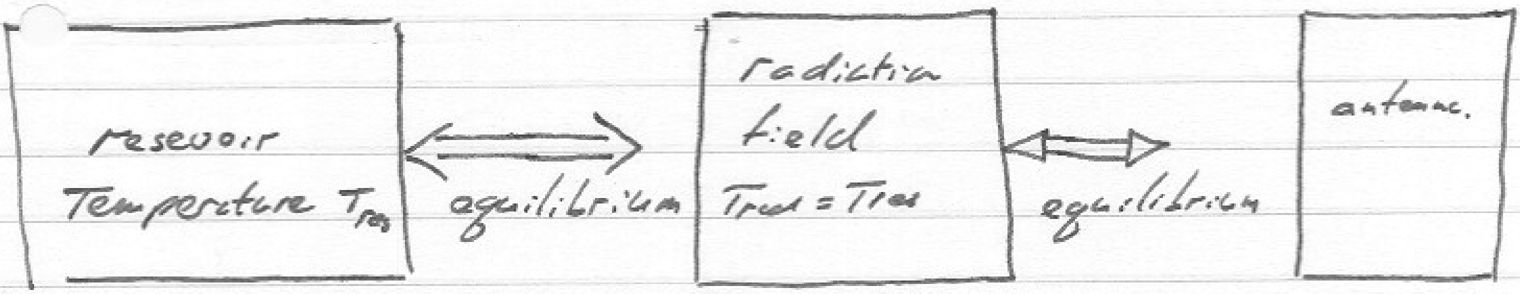
$\therefore$ E_{n1} and E_{n2} are clearly correllated!

Antenna Noise

The terminal impedance of an antenna is modeled as so:



If we couple the antenna to a radiation field, we can compute its available noise power:



So, the antenna has available noise power

$\simeq kT_{rad}$. $\Rightarrow \left[\frac{\partial \langle P_\omega \rangle}{\partial f} = \frac{1}{2} hf + \frac{hf}{\exp(hf/kT) - 1} \right]$

Note that T_{rad} is the temperature of the radiation field to which the antenna is coupled.

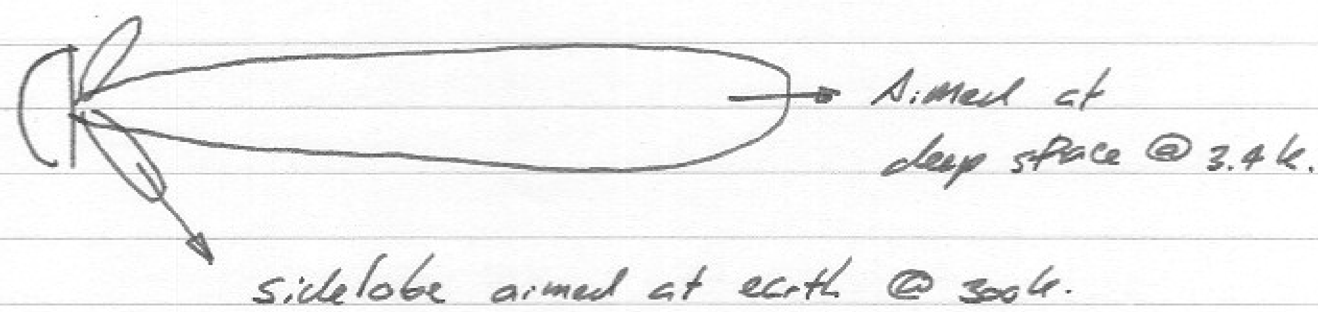
This depends very much on where the antenna is aimed.

Aimed at deep space:

- Cosmic background radiation: 3.4 K.

Aimed near planet (earths) surface: ~~4~~ 300 K:

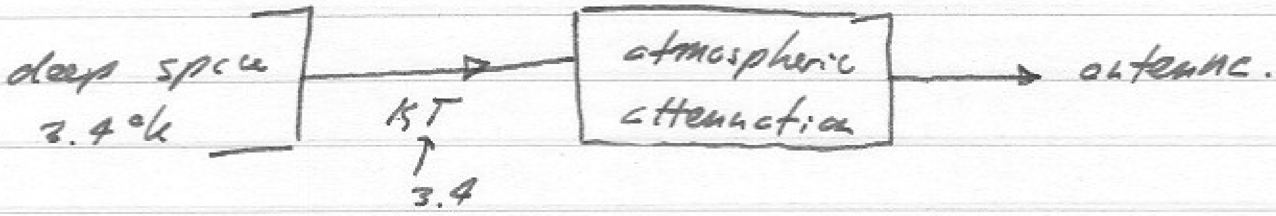
Antenna's for radio astronomy must be designed for very low sidelobes:

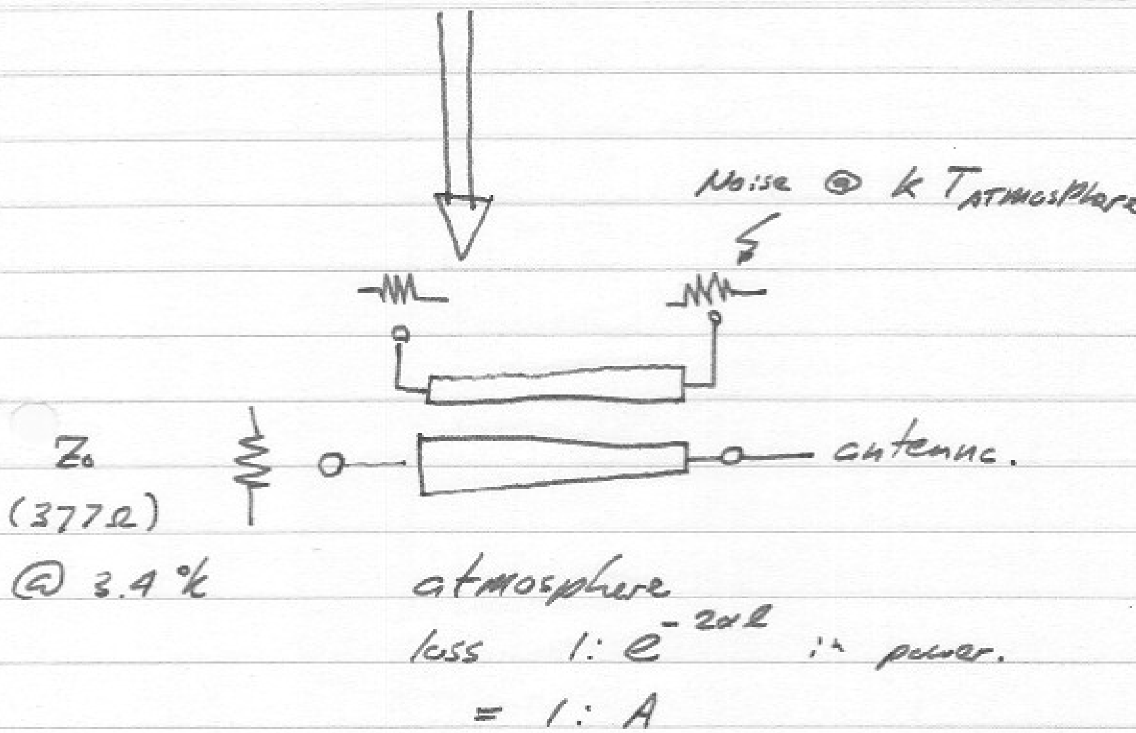


[A 1% sidelobe aimed at earth generates as much noise power as the target!

Antennas for Astronomy are made ~~as~~ big as much for sidelobe noise suppression as they are for gain or directivity

Atmospheric Effects can be modelled as an attenuator:

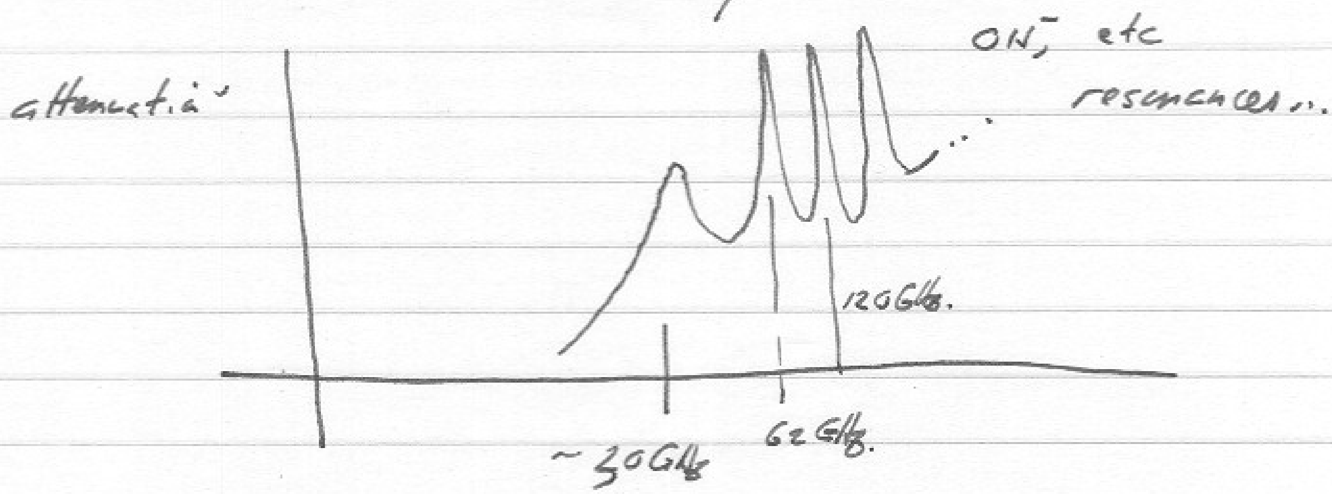




the antenna will see an available noise power

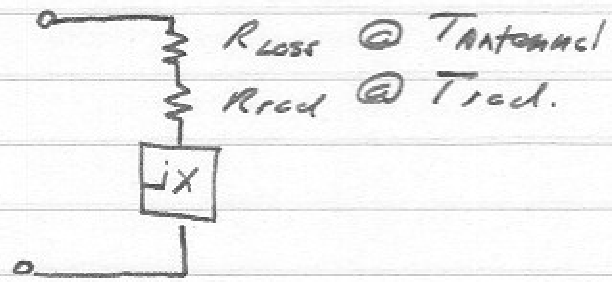
$$T_{rad} \cdot (A) + T_{ATMOSPHERE} \cdot (1 - A)$$

The atmosphere is at 250 - 300°K, and has an attenuation curve somethingk like this



So an antenna pointed at the Zenith would produce strong noise peaks at 30, 62, & 120 GHz.

Real antennas also have conductor resistance, from which the model becomes:



As deep space / satellite antennas see $T_{rad} << 300\,°K$, it is imperative to obtain $R_{loss} << R_{rad}$.

Note that terrestrial communications systems will see $T_{rad} \sim 300\,°K$, whilst satellite receivers ~~see~~ may see $T_{rad} \sim 10 - 50$ K. Requirements will differ greatly.

Note that noise figure:

$$F \triangleq \frac{S/N|_{IN}}{S/N|_{out}} = \frac{kT + N_{AMP}}{kT}$$

... where N_{AMP} is the amplifiers input-referred noise power contribution. F is normally specified for a 300K background for kT, yet an amplifier may be used with an antenna having $T_{rad} << 300°K$. In this case, $F=1.1$ vs $F=1.01$ can represent vastly different % noise contributions from the amplifier. Better to either use F defined at the antenna ~~Noise~~ radiation noise temperature or use the equivalent noise temperature:

$$\boxed{T_{eq} = N_{AMP}/k.}$$

From which: $$\boxed{T_{eq} = [F(T_0) - 1] \cdot T_0}$$

where $F(T_0)$ is the noise figure at reference temperature T_0

or $$\boxed{F(T_0) = 1 + \frac{T_{eq}}{T_0}}$$

note that noise temperature is <u>not</u> <u>physical</u>

a 10dB 50Ω attenuator at 300°K room temperature has a 3000 K noise temperature. There is nothing that hot in the thermodynamic sense.

We can write the Friis Noise formula in terms of noise temperatures:

$$F_T = F_1 + \frac{F_2 - 1}{G_{av1}} + \frac{F_3 - 1}{G_{av1} G_{av2}} + \cdots$$

$$T_0 + T_{eqT} = T_0 + T_{eq1} + \frac{T_{eq2}}{G_{av1}} + \frac{T_{eq3}}{G_{av1} G_{av2}} + \cdots$$

$$\boxed{T_{eqT} = T_{eq1} + \frac{T_{eq2}}{G_{av1}} + \frac{T_{eq3}}{G_{av1} G_{av2}} + \cdots}$$

The T_{eq}'s are calculated for each device assuming the output impedance of the prior stage. This should remind you how little T_{eq} has to do with a physical temperature $\left[\frac{1}{T} \triangleq \frac{\partial S}{\partial E} \right]$