

# Effect of on Resistance Modulation in RF Switches Linearity

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**Abstract** Radio frequency (RF) switch modeling has evolved as semiconductor processes have been developed and refined. However, to date no modeling has included the effect of the variation of the resistance of a switch on its linearity when the switch is in the on state. This paper presents an analysis that predicts the effect on signal linearity when the series resistance is modulated.

**Keywords** FET switch, Linearity, pHEMT switch, RF switch

## 1. Introduction

Field Effect Transistor (FET) RF switches are used in nearly all wireless communication devices, from cell phones to Wi-Fi to GPS systems [1-9]. With a growing communications market, their numbers will also grow. However, the continued growth is contingent on the constant evolution and refinement of RF switches. Specifically, newer communications architectures and implementations for fourth generation communications (4G), such as 3GPP Long Term Evolution (LTE) and IEEE802.11n, require higher frequencies and corresponding bandwidths while minimizing the distortion [1, 5, 7-9]. Minimizing distortion translates to maximizing linearization. There have been many steps over the past several years to improve the linearity of switches to meet the demanding specifications. Correspondingly, defining and modeling FET as switches have also improved [12-18], although most work has been done viewing the FET as simply a microwave device or as an amplifier [19-34]. This author has found no analysis specifically tying modulation of the ON resistance ( $R_{on}$ ) to linearity or what inherent characteristics in a FET, such as transconductance, might lead to ON resistance sensitivity to modulation.

## 2. Background

The wireless performance requirements in today's RF

devices call out very stringent IP3 specifications depending upon the architecture and options used (65 dBm for GSM and WCDMA) – and have been very difficult to meet [35 – 37]. Nonetheless, researchers and companies have striven to create switches to meet these requirements. In 2006, Mil'shtein and Liessner et. al. introduced methods to improve linearity of FET switches [38-40]. In 2010, Andrew Dearn and Liam Devlin of Plextek Ltd. presented high linearity SPDT switches with IP3 at 1.9 GHz up to 54 dBm, with the second harmonic recorded at 80 dBc for that device [41]. Also, in 2010, Michael Yore, et.al. of Triquint Semiconductor reported a SP7T pHEMT switch with IMD2 typically better than -100 dBm and IMD3 typically better than -105dBm [42]. As of 2013, commercially, Skyworks currently offers an ultra-high linearity SP2T switch, SKY13405-409LF, capable of a typical IIP3 of 68 dBm [37].

## 3. Analysis

Looking at the DC I-V characteristics of a FET switch, the channel resistance for a switch when it is in the ON state is effectively the inverse slope of the  $I_{ds}$  versus  $V_{ds}$  for a given  $V_{gs}$  at low values of  $V_{ds}$ :

$$R_{on} = \frac{1}{\frac{I_{ds}}{V_{ds}}} = \frac{V_{ds}}{I_{ds}} \quad (1)$$

In this linear region of  $I_{ds}$  versus  $V_{ds}$ , this ratio is essentially a constant with a constant  $V_{gs}$ . However, this ratio – or equivalently the resistance,  $R_{on}$ , - changes as a function of  $V_{gs}$ . Since  $R_{on}$  is a function of  $V_{gs}$ , taking the derivative of (1) with respect to  $V_{gs}$  will look as follows:

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$$\frac{\partial R_{on}}{\partial V_{gs}} = \frac{\frac{\partial V_{ds}}{\partial V_{gs}}}{\frac{\partial I_{ds}}{\partial V_{gs}}} = \frac{1}{I_{ds}} \left( \frac{\partial V_{ds}}{\partial V_{gs}} \right) - \frac{V_{ds}}{I_{ds}^2} \left( \frac{\partial I_{ds}}{\partial V_{gs}} \right) \quad (2)$$

$$\frac{\partial R_{on}}{\partial V_{gs}} = -\frac{1}{I_{ds}} \left( \frac{V_{ds}}{I_{ds}} \right) \left( \frac{\partial I_{ds}}{\partial V_{gs}} \right) \quad (3)$$

Reviewing the definition of the transconductance,  $g_m$  (4), and substituting it into equation (3), along with equation for  $R_{on}$  (1), we establish a relationship between the sensitivity of  $R_{on}$  to  $V_{gs}$ ,  $V_{ds}$ , and  $I_{ds}$  in equation (5).

$$g_m = \left( \frac{\partial I_{ds}}{\partial V_{gs}} \right) \quad (4)$$

$$\frac{\partial R_{on}}{\partial V_{gs}} = -\frac{R_{on} \cdot g_m}{I_{ds}} \quad (5)$$

Equation (5) establishes a direct correlation between the change in  $R_{on}$  to its dc value, the value of the dc current ( $I_{ds}$ ), and the transconductance ( $g_m$ ), to the variation in the gate voltage ( $\partial V_{gs}$ ). It is seen that the smaller  $R_{on}$  is, the smaller the sensitivity to change in  $V_{gs}$ . In addition, as the dc  $I_{ds}$  increases, the variation in  $R_{on}$  also diminishes. This is because a larger  $I_{ds}$  implies a smaller  $R_{on}$ ; and the smaller  $R_{on}$  is, the smaller the impact that any variation on it has on the overall signal.

Now we need to relate this derivation to switch linearity for a series switch, and to the 3<sup>rd</sup> intermodulation products in particular. In Figure 1 is shown the simple power transfer circuit of a switch in the ON state.  $P_{in}$  and  $P_{out}$  are described in equations (6) and (7). Manipulating equations (6), (7), and (8), and realizing that  $R_{on}$  is typically much less than the  $Z_L$ , we can derive the relationship of  $P_{in}$  to  $P_{out}$  relative to  $R_{on}$ . The associated insertion loss is described by equation (10).

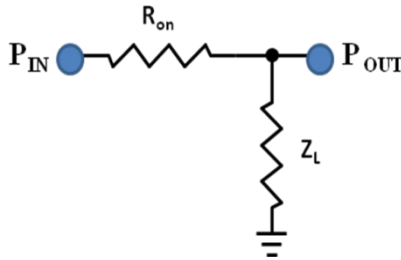


Figure 1. Switch ON state with load

$$P_{in} = \frac{V_{in}^2}{Z_L + R_{on}} \quad (6)$$

$$P_{out} = \frac{V_{out}^2}{Z_L} \quad (7)$$

$$V_{out} = V_{in} \frac{Z_L}{Z_L + R_{on}} \quad (8)$$

$$P_{out} \approx P_{in} \cdot \left[ \frac{1}{1 + \frac{2R_{on}}{Z_L}} \right] \quad (9)$$

$$InsertionLoss \approx 10 \log \left[ \frac{1}{1 + \frac{2R_{on}}{Z_L}} \right] \quad (10)$$

Using a Taylor Series expansion [43], a weakly nonlinear response can be modeled around a quiescent point, such as a switch either in the ON state or OFF state:

$$x(t) = A \cos(\omega_1 t) + A \cos(\omega_2 t) \quad (11)$$

$$y(t) = G(t) \cdot A \cos(\omega t + \varphi(\omega)) \quad (12)$$

$$y_N(t) = G_1 \cdot x(t) + G_2 \cdot x(t)^2 + G_3 \cdot x(t)^3 + \dots$$

$$G_n = \frac{1}{n!} \cdot \frac{d^n G}{dx^n} \quad (13)$$

Deriving for a two-tone solution, as used for IP3 measurements, equation (13) becomes as follows:

$$y_{N,1}(t) = G_1 \cdot A \cos(\omega t - \varphi_1) + \frac{G_2 \cdot A^2}{2} + \frac{G_2 \cdot A^2}{2} \cos^2(2\omega t - 2\varphi_2) + \frac{3G_3 \cdot A^3}{4} \cos^3(\omega t - \varphi_3) + \frac{3G_3 \cdot A^3}{4} \cos^3(3\omega t - 3\varphi_3) + \dots \quad (14)$$

Now we can proceed with deriving the appropriate IM3 calculation. From equation (9) and (14), we can separate out the IM3 products:

$$IM3 = \frac{3G_3 \cdot A^3}{4} \quad (15)$$

If  $P_{in}$  is a constant value  $A$ , and we used the simplified attenuation factor in (9) we can now define the  $G$  function from (15) as:

$$G = \left[ \frac{1}{1 + \frac{2R_{on}}{Z_L}} \right] \quad (16)$$

To find  $G_3$ , from equation (13) we will use equation (17), and identifying  $\partial R_{on}$  and  $\partial V_{gs}$ , and calling  $k$  the RF coupling, we derive equation (20).

$$G_3 = \frac{1}{3!} \cdot \frac{d^3 G}{dx^3} = \frac{1}{6} \cdot \frac{d^3 G}{dx^3} \quad (17)$$

$$\partial R_{on} = -\frac{R_{on} g_m}{I_{ds}} \cdot \partial V_{gs} \quad (18)$$

$$\partial V_{gs} = k A \cos(\omega t) \quad (19)$$

$$\frac{d^3 G(t)}{dt^3} = -\left[ \left( \frac{k R_{on} g_m}{Z_L I_{ds}} \right)^3 \cdot \frac{8}{\delta^4} \right] \quad (20)$$

Where  $\delta$  is defined as:

$$\delta = \left( 1 + \frac{2 R_{on}}{Z_L} - \frac{A \sqrt{2} k R_{on} g_m}{Z_L I_{ds}} \right) \quad (21)$$

Plugging (20) into (15) yields the final equation for finding the amplitude of 3<sup>rd</sup> order intermodulation product with a modulated  $R_{on}$ . The logarithmic version of IM3 in dB is found in equation (23).

$$IM3 = \left| -\left[ \left( \frac{k R_{on} g_m}{Z_L I_{ds}} \right)^3 \cdot \frac{6}{\delta^4} \right] \cdot A^3 \right| \quad (22)$$

$$IM3_{dB} = 20 \log \left| -\left[ \left( \frac{k R_{on} g_m}{Z_L I_{ds}} \right)^3 \cdot \frac{6}{\delta^4} \right] \cdot A^3 \right| \quad (23)$$

To calculate the IIP3, first  $P_{in}$  must be properly defined. Since  $\partial R_{on}$  is much less than  $Z_L$ ,

$$P_{in} = \frac{A^2}{2(Z_L + R_{on})} \quad (24)$$

Since measurements are made using dBm as the standard, the amplitude is multiplied by a factor of 1000 to properly scale.

$$P_{in, dBm} = 10 \log \left( \frac{A^2 \cdot 500}{Z_L + R_{on}} \right) \quad (25)$$

And the IIP3 is shown below:

$$IIP3 = \frac{3}{2} P_{in} - 10 \log \left| -\left[ \left( \frac{k R_{on} g_m}{Z_L I_{ds}} \right)^3 \cdot \frac{6}{\delta^4} \right] \cdot A^3 \right| \quad (26)$$

However, note that  $I_{ds}$  is not arbitrary – it is tied to the power input (or, equivalently, the RMS input voltage), the ON resistance, and the load impedance as follows:

$$I_{ds} = \sqrt{\frac{P_{in}}{Z_L + R_{on} + \partial R_{on}}} \approx \sqrt{\frac{P_{in}}{Z_L + R_{on}}} \quad (27)$$

And therefore:

$$IIP3 = \frac{3}{2} P_{in} - 10 \log \left| -\left[ \left( \frac{\sqrt{8} k R_{on} g_m \sqrt{Z_L + R_{on}}}{\sqrt{Z_L}} \right)^3 \cdot \frac{6}{\delta^4} \right] \right| \quad (28)$$

With  $\delta$  represented now as:

$$\delta = \left( 1 + \frac{2 R_{on}}{Z_L} - \frac{2 k R_{on} g_m (Z_L + R_{on})}{Z_L} \right) \quad (29)$$

And as expected, the power cancels out in this equation since compression and dispersive effects are not in the model. This leaves the IIP3 a function of  $R_{on}$ ,  $g_m$ ,  $Z_L$ , and the coupling factor  $k$ . Note that this analysis only shows the affect and sensitivity of a switch due to transconductance in the region of interest and fluctuations to  $R_{on}$ ; it does not provide a model for a specific PHEMT structure otherwise.

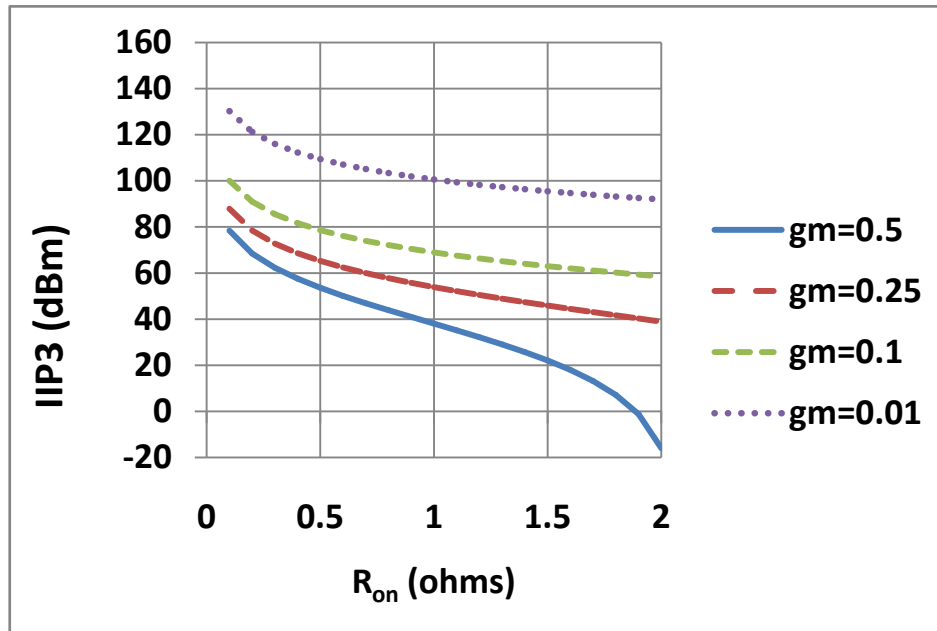
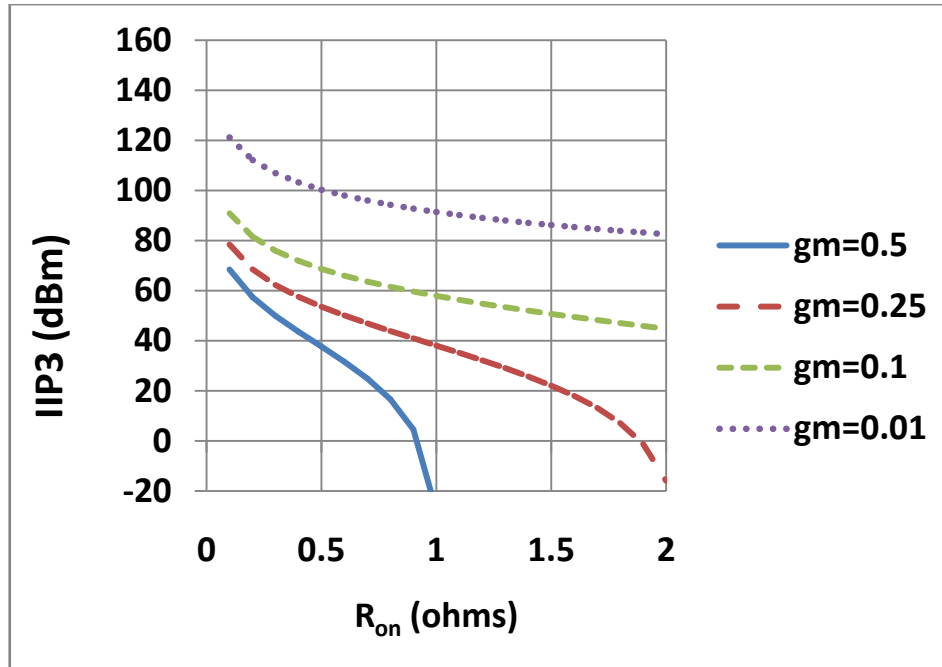
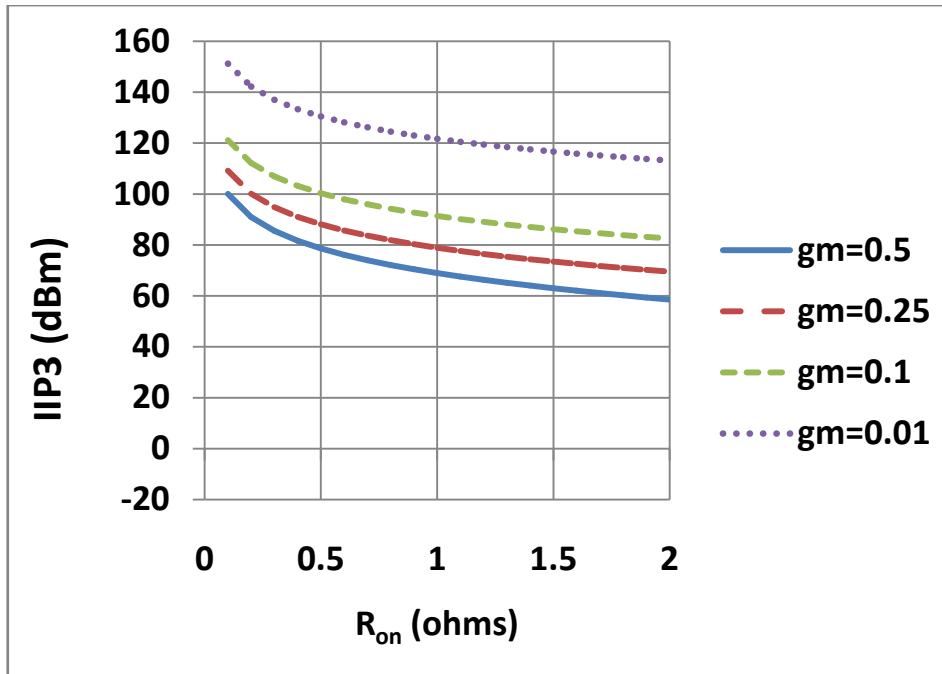


Figure 2. IIP3:  $k=0.5$

Figure 3. IIP3:  $k=1.0$ Figure 4. IIP3:  $k=0.1$ 

## 4. Results

A series of plots were made to show the predicted impact on switch linearity.  $R_{on}$  was swept over a range of  $0.1\Omega$  to  $2\Omega$ , and four different transconductance values were selected: 500mS, 250mS, 100mS, and 10mS. The RF coupling factor was selected to be at 50%, 100%, and 10% in three separate plots. The load impedance was set to  $50\Omega$ . Figures 2, 3, and 4 show the results. Figures 5 and 6 display the affect of a 2:1 VSWR on the load, with Figure 5 calculated for  $100\Omega$  and Figure 6 determined for a  $25\Omega$ . Note that the diminished  $Z_L$

also represents potential loading caused by an OFF switch element on the switch.

## 5. Discussion

Looking at the linearity for the ON state of RF FET switches, analysis and calculations suggest the hypothesis of a sharp transconductance profile for optimum linearity. This reduces the affect of RF coupling onto the gate of the FET. Using the equations established, the plots in figures 2 – 4

show the clear advantage to minimizing the  $g_m$  value in a switch FET when it is in the ON state. Higher  $g_m$  in the ON state produces more distortion at the output, even with minimal RF coupling. Referring to figure 4, which shows results based on a coupling factor of  $k = 0.1$ , a device with a  $R_{on} = 1$  ohm and a  $g_m = 500\text{mS}$  can only achieve an IIP3 of 70dBm, whereas a device with the same  $R_{on}$  but with a  $g_m = 100\text{mS}$  can reach an IIP3 = 92dBm. This is a 22dB improvement. As the coupling factor increases to 0.5 and 1.0 as shown in figures 2 and 3 respectively, the improvement achieved by the lower  $g_m$  increases even further: at  $k = 0.5$  there is a 31dB gain, and at  $k = 1.0$  there is a >80dB improvement. Despite that, a high transconductance is

preferred for the transition of switch states and higher  $I_{ds}$ . A sharp, peaked, impulse-shaped transconductance profile is suggested to be best, since it allows for a lower  $g_m$  during a switch ON state. The coupling factor,  $k$ , affects the intermodulation products as expected - more coupling results in less linearity and poorer IP3 values. While this occurs regardless of  $g_m$ , it is demonstrated in figures 2 -6 and discussed above, that a lower  $g_m$  compared to a higher  $g_m$  improves the IP3 performance. It is noticed in Figures 5 and 6, that a lesser load impedance actually improves the linearity. The reason for this is that the higher  $I_{ds}$  is, the greater the linearity; the compression and dispersion effects are not modeled here.

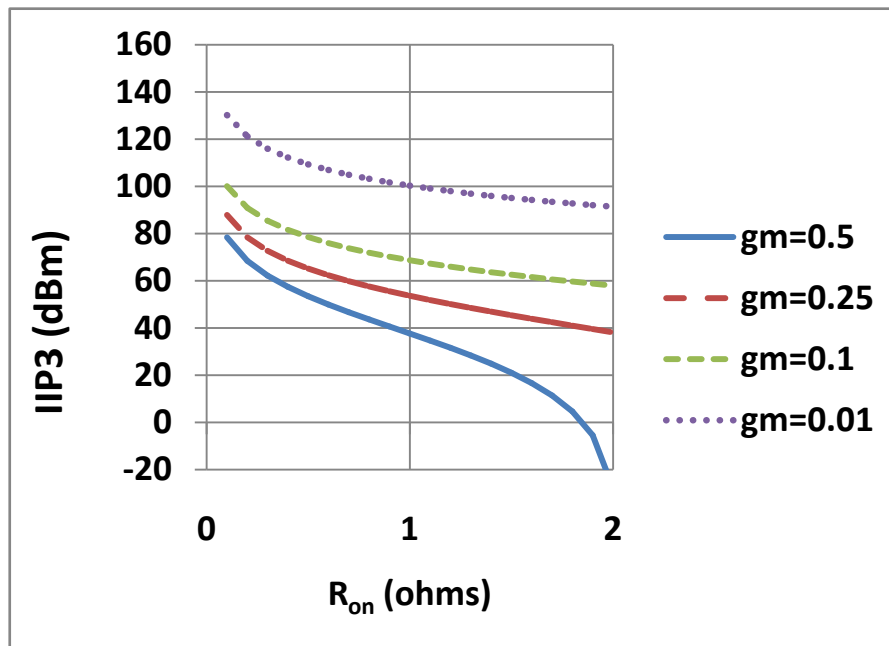


Figure 5. IIP3:  $k=0.5$ ,  $Z_L = 100$  ohms

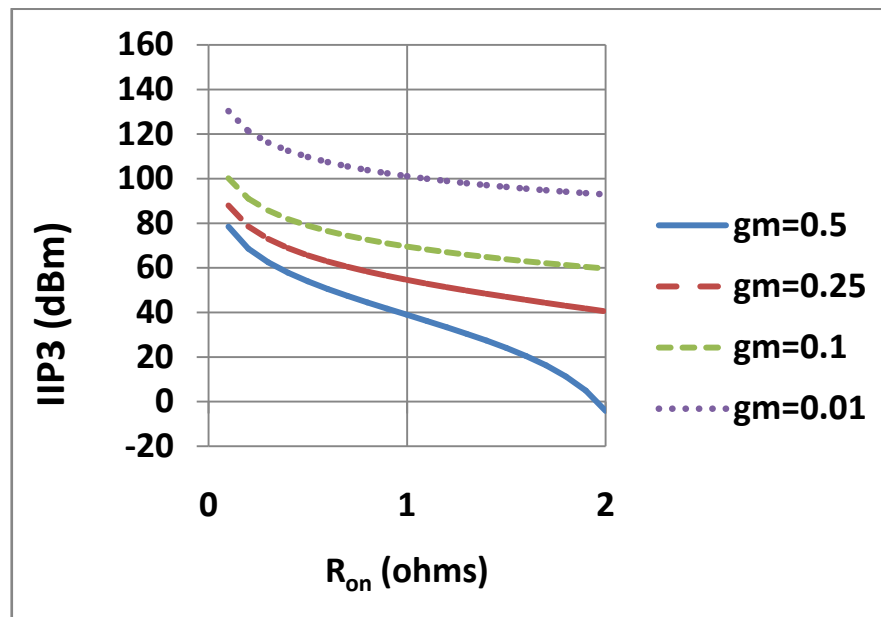


Figure 6. IIP3:  $k=0.5$ ,  $Z_L = 25$  ohms

## 6. Conclusions

After extensive analysis, a new approach to characterizing FET switches has been proposed, analyzed, and mathematically supported. A new analytical tool has been created to provide guidance to designing switches with high linearity, and it has been shown that a lower transconductance value when the FET is in the ON state yields superior linearity over a FET with higher transconductance, suggesting a different transconductance profile than what is often proposed, such as an impulse.

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