EM field analysis for phase cancellation in backscattering tag to tag systems

Jinghui Jian

1 Objective

In this document, we will analyze the phase cancellation problem from electromagnetic perspective, and propose solving this problem by apply enhanced multi-phase backscatter modulation.

2 Backscattered field of tag

In this section, we are going to introduce some antenna fundamental first. Most of the fundamental theory is coming from the antenna book [1]. The far-zone electric field $\mathbf{E}_a$ observed at distance d can be expressed by its driven current I_a and antenna equivalent vector effective length $\mathbf{l}_e$:

$$\mathbf{E}_a = -j\eta \frac{\beta I_a}{4\pi d}\mathbf{l}_e e^{-j\beta d} = -j\mathbf{i}_{\mathbf{l}_e}\frac{\eta I_a}{2\lambda d}\left|\mathbf{l}_e\right| e^{-j\beta d}, \tag{1}$$

where $\mathbf{i}_{\mathbf{l}_e}$ is the unit vector of antenna effective length, η is the medium impedance which is 377 Ω for free space and air, β is the propagation constant given by $2\pi/\lambda$, and λ is the wave length. According to [2], $\mathbf{l}_e$ is associated with the antenna gain G and radiation resistance R_A of the antenna, which is the real part of Z_A:

$$|\mathbf{l}_e| = \sqrt{\frac{G\lambda^2 R_A}{\pi\eta}}. \tag{2}$$

Effective length $\mathbf{l}_e$ can also be used to derive the induced open-circuit voltage V_{op} when incident electric field for receiving antenna is $\mathbf{E}_{inc}$:

$$V_{op} = \mathbf{E}_{inc} \cdot \mathbf{l}_e. \tag{3}$$

The Thévenin equivalent circuits for transmitting antenna and receiving antenna are shown in Fig. 1 and Fig. 2 respectively. When impedances of transmit and receive antennas are conjugate-matched to transmitter and receiver, it's very easy to derive the power relationship of a general radio link consisting a transmitter and a receiver:

$$P_{\mathrm{r}} = \frac{P_{\mathrm{t}} G_{\mathrm{t}} G_{\mathrm{t}} \lambda^2}{(4\pi d)^2}, \tag{4}$$

where P_{t} and P_{r} are the general transmitting and receiving power in the link, G_{t} and G_{r} are corresponding transmitter and receiver antenna gain, d is the communication distance and λ is the wavelength. And this euqation is the famous Friis's formula [1].

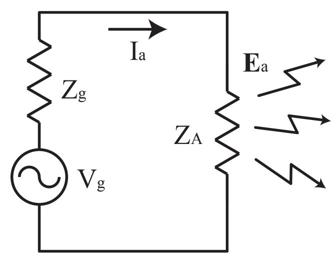

Figure 1: Thévenin equivalent circuits for transmitting antenna and radiation field

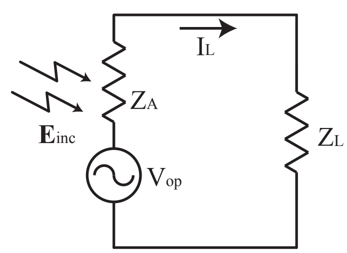

Figure 2: Thévenin equivalent circuits for receiving antenna

Based on antenna scattering theory developed by Hansen [2], the scattered field from an antenna in terms of load at the antenna Z_L as:

$$\mathbf{E}_{scat}(Z_L) = \mathbf{E}_{scat}(Z_A^*) + \Gamma_L^* \frac{\mathbf{E}_a I_m^*}{I_a}. \tag{5}$$

The first term $\mathbf{E}_{scat}(Z_A^*)$ is the structural scattering field cased by the current I_m^* induced on the antenna surface by incident field when the load of antenna is conjugate matched; it is independent of load impedance Z_L. The second term $\mathbf{E}_a$ is the antenna mode scattering field related with antenna radiation characteristic, in which $\mathbf{E}_a$ is the radiation field when I_a is the driven current, and Γ_L^* is conjugate antenna reflection coefficient at different modulation states, given by

$$\Gamma_{L,i}^* = \frac{Z_A^* - Z_{L,i}}{Z_{L,i} + Z_A}, i = 1, 2. \tag{6}$$

Eq. (5) can be easily transferred into the following form [3]:

$$\mathbf{E}_{scat}(Z_L) = \frac{\mathbf{E}_a I_m^*}{I_a} (A + \Gamma_L^*), \tag{7}$$

then A is expressed as:

$$A = \frac{\mathbf{E}_{scat}(Z_A^*) I_a}{\mathbf{E}_a I_m^*}. \tag{8}$$

A is a complex, the induced antenna surface current related coefficient, which is dependent on antenna geometrical layout and EM properties of the material. For thin layer dipole-like antenna, A can be approximated as 1 [1].

For conjugate matched antenna, the corresponding induced conjugate current I_m^* is

$$I_m^* = \frac{V_{op}}{Z_A + Z_A^*} = \frac{\mathbf{E}_{inc} \cdot \mathbf{l}_e}{2R_A}. \tag{9}$$

After substituting I_m^* into Eq. (7), and applying the radiation equation Eq. (1), the scattered field observed at d distance away from backscattering antenna is

$$\mathbf{E}_{scat}(Z_L) = -j \frac{\eta e^{-j\beta d}}{4\lambda R_A d} (\mathbf{E}_{inc} \cdot \mathbf{l}_e)\, \mathbf{l}_e\, (1 + \Gamma_L^*) \tag{10}$$

$$= -j \mathbf{i}_{\mathbf{l}_e} \sqrt{p_e} \frac{\eta e^{-j(\beta d + \theta_{\mathbf{E}_{inc}})}}{4\lambda R_A d} |\mathbf{E}_{inc}|\, |\mathbf{l}_e|^2 (1 + \Gamma_L^*), \tag{11}$$

where $\theta_{\mathbf{E}_{inc}}$ is the phase change introduced by incident field $\mathbf{E}_{inc}$, p_e polarization efficiency:

$$p_e = \frac{|\mathbf{E}_{inc} \cdot \mathbf{l}_e|^2}{|\mathbf{E}_{inc}|^2 |\mathbf{l}_e|^2}; \tag{12}$$

p_e equals to 1 for polarization match. Relating $\mathbf{l}_e$ with antenna gain G and resistnace Z_A by (2), $\mathbf{E}_{scat}(Z_L)$ can be further transferred to

$$\mathbf{E}_{scat}(Z_L) = -j \mathbf{i}_{\mathbf{l}_e} \sqrt{p_e} \frac{G\lambda}{4\pi d} e^{-j(\beta d + \theta_{\mathbf{E}_{inc}})} |\mathbf{E}_{inc}| (1 + \Gamma_L^*), \tag{13}$$

If we apply the above derived equations into the RCS definition formula:

$$\sigma = \lim_{d \to \infty} \left[4\pi d^2 \frac{|\mathbf{E}_{scat}|^2}{|\mathbf{E}_{inc}|^2} \right], \tag{14}$$

where d is the observation distance, we can get differential RCS as shown in Eq. (**??**).

3 The phase cancellation problem

Fig. 3 shows a BBTT system with one CW exciter and two tags, one is labeled as Tx tag to backscatter the message to the other tag labeled as Rx tag. The respective distances between three devices are shown in the figure. The CW exciter has an antenna with gain G_{CW}, resistance R_{ACW}, the two tags have identical antenna characteristic gain G_{tag} and resistance R_{Atag}. Here we assume that all the antennas in the system have been positioned optimally so that there is no polarization losses, which means the vector effective length of all the antenna have the same direction $\mathbf{i}_{\mathbf{l}_e}$ and $p_e = 1$. By

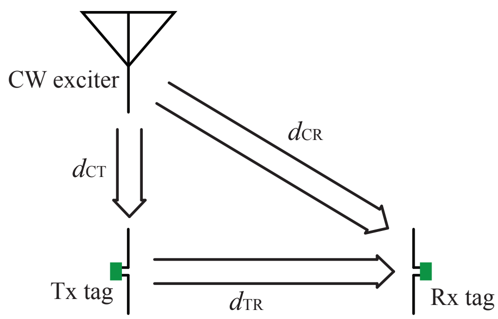

Figure 3: A BBTT system with one CW exciter and two tags

using Eq. (1) and (2), we can get the electrical field received by Tx and Rx tags from exciter as:

$$\mathbf{E}_{\mathrm{tC}} = \mathbf{i}_{\mathbf{l}_e} I_{\mathrm{CW}} \sqrt{\frac{\eta G_{\mathrm{CW}} R_{\mathrm{ACW}}}{4\pi d_{\mathrm{CT}}^2}} e^{-j(\beta d_{\mathrm{CT}} + \frac{\pi}{2})} = \mathbf{i}_{\mathbf{l}_e} A_{\mathrm{tC}} e^{j\theta_{\mathrm{tC}}}, \tag{15}$$

$$\mathbf{E}_{\mathrm{rC}} = \mathbf{i}_{\mathbf{l}_e} I_{\mathrm{CW}} \sqrt{\frac{\eta G_{\mathrm{CW}} R_{\mathrm{ACW}}}{4\pi d_{\mathrm{CR}}^2}} e^{-j(\beta d_{\mathrm{CR}} + \frac{\pi}{2})} = \mathbf{i}_{\mathbf{l}_e} A_{\mathrm{rC}} e^{j\theta_{\mathrm{rC}}}, \tag{16}$$

where I_{CW} is driving current in the CW exciter. Then from Eq. (13), we can get the backscattered signal received by Rx tag from Tx tag at two different modulation states as:

$$\begin{aligned} \mathbf{E}_{\mathrm{rxBck,i}} &= \mathbf{i}_{\mathbf{l}_e} \frac{G_{\mathrm{tag}} \lambda}{4\pi d_{\mathrm{TR}}} e^{-j(\beta d_{\mathrm{TR}} + \frac{\pi}{2})} \mathrm{E}_{\mathrm{tC}} \left(1 + \Gamma_{\mathrm{L,i}}^*\right) \\ &= \mathbf{i}_{\mathbf{l}_e} A_{\mathrm{sconj}} e^{j\theta_{\mathrm{sconj}}} \left(1 + \rho_{\mathrm{L,i}}^* e^{j\theta_{\mathrm{L,i}}^*}\right), \mathrm{i} = 1, 2 \end{aligned} \tag{17}$$

where

$$A_{\mathrm{sconj}} = \frac{G_{\mathrm{tag}} \lambda}{4\pi d_{\mathrm{TR}}} A_{\mathrm{tC}}, \tag{18}$$

$$\theta_{\mathrm{sconj}} = \theta_{\mathrm{tC}} - (\beta d_{\mathrm{TR}} + \frac{\pi}{2}), \tag{19}$$

$$\rho_{\mathrm{L,i}}^* = \left|\Gamma_{\mathrm{L,i}}^*\right|, 0 \le \rho_{\mathrm{L,i}}^* \le 1 \tag{20}$$

$$\theta_{\mathrm{L,i}}^* = \angle \Gamma_{\mathrm{L,i}}^*. \tag{21}$$

So without considering the other multipath and environmental field, the total electric field received by Rx tag is:

$$\mathbf{E}_{\mathrm{rxTot,i}} = \mathbf{i}_{\mathrm{l}_e}\left[A_{\mathrm{rC}}e^{j\theta_{\mathrm{rC}}} + A_{\mathrm{sconj}}e^{j\theta_{\mathrm{sconj}}} + A_{\mathrm{sconj}}\rho^*_{\mathrm{L,i}}e^{j\left(\theta_{\mathrm{sconj}}+\theta^*_{\mathrm{L,i}}\right)}\right] \tag{22}$$

$$= \mathbf{i}_{\mathrm{l}_e}\left(E_{\mathrm{CW}} + E_{\mathrm{sconj}} + \Gamma^*_{\mathrm{L,i}}E_{\mathrm{sconj}}\right) \tag{23}$$

$$= \mathbf{i}_{\mathrm{l}_e}\left(E_{\mathrm{CW}} + E_{\mathrm{sconj}} + E_{\mathrm{mod,i}}\right) \tag{24}$$

$$= \mathbf{i}_{\mathrm{l}_e}A_{\mathrm{rxTot,i}}e^{j\theta_{\mathrm{rxTot,i}}}\mathrm{i} = 1, 2. \tag{25}$$

In Eq. (24), the first two terms are CW signal and conjugate-matched structural scattering field, which are stationary fields during two modulation states when all other factors are fixed; the third term is the modulation signal, the change of which will cause the change of total field at receiver side. The phasor form of signal superposition is shown in Fig. 4, the angle shift of smith chart can caused by phase shifting of transmission line and imaginary impedance of antenna. By either modulating the amplitude or phase of conjugate antenna reflection coefficient $\Gamma^*_{L,i}$, we can change the phase and amplitude of total signal seen by the receiver: the first modulation method is ASK modulation, and the second one is called PSK modulation. The envelop detector demodulate the signal by detecting the power difference between the total signal at two states.

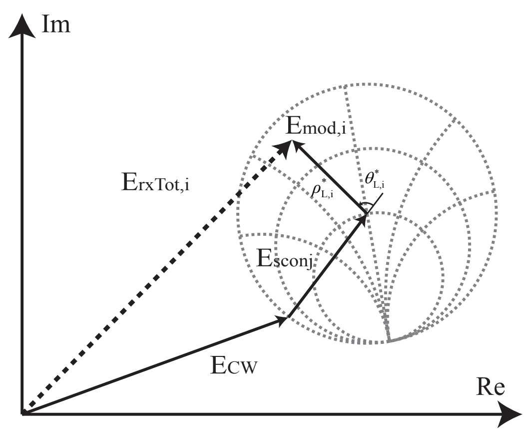

Figure 4: Phasor form of total received signal

However, any change on the environmental factors, like the relative positions of the three devices, radiation power of the CW signal and etc, will cause the change of first two “stationary” terms, which may lead the combination of fields to the positions that even $\Gamma^*_{L,i}$ modulates between two states, the amplitudes of total received power still remain constant. Then the envelop detector is not able to detect the difference of the signals. We call this situation as phase cancellation. Fig. 5 and 6 show the cases ASK modulation without and with the phase cancellation. PSK modulation also has the same problem, and the phasor illustration is shown in Fig. 7 and Fig. 8.

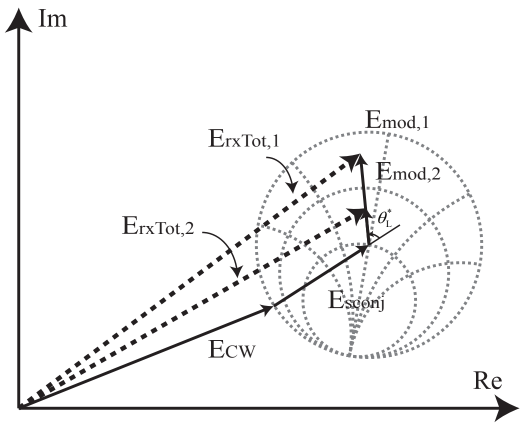

Figure 5: A phasor diagram of ASK modulation when there is no phase cancellation

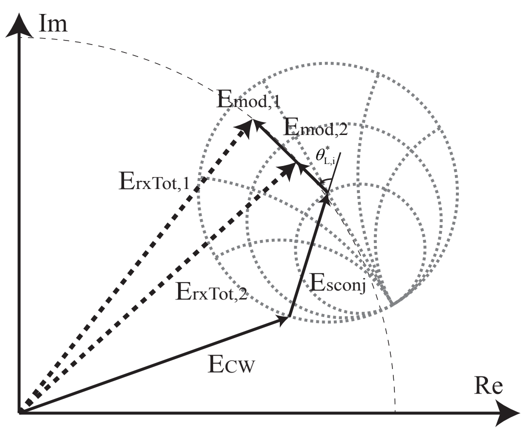

Figure 6: A phasor diagram of ASK modulation when there is phase cancellation

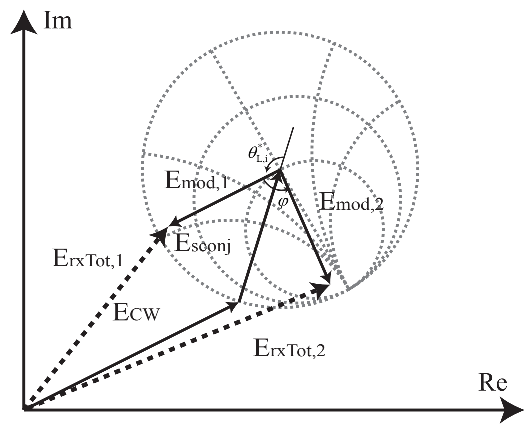

Figure 7: A phasor diagram of PSK modulation when there is no phase cancellation

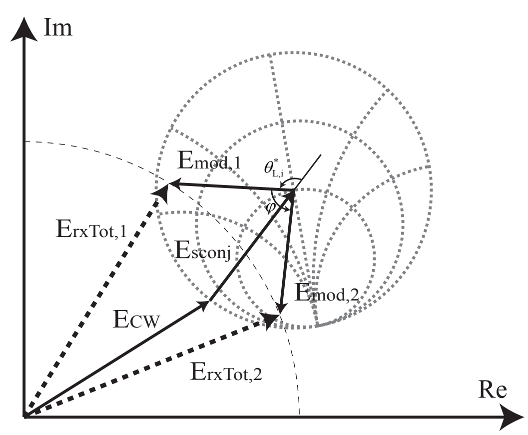

Figure 8: A phasor diagram of PSK modulation when there is phase cancellation

4 Enhanced multi-phase backscatter modulator design

To solve phase cancellation problem, some paper propose applying multiple antennas [4] to extract EM fields at different positions for different phases; the combination of the signal will appear power-difference to the detector. This is an effective method but will make the tag kind of bulky. Also when the CW frequency is low, the distance between the antennas need to be even longer to get enough phase difference. Our proposed solution is applying enhanced backscatter modulator with phase diversity introduced. The tag will backscatter the information through two different phase channel in two successive time slots with a deterministic phase difference θ_n between the two channels; then if there is phase cancellation happen in one channel, the other channel with phase shifting θ_n will avoid it as illustrated in Fig. 9.

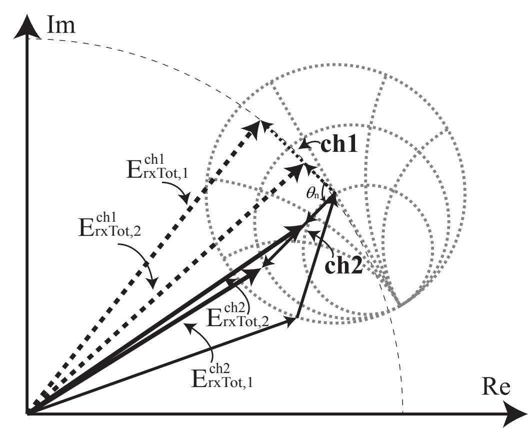

Figure 9: Solving phase cancellation problem by backscattering signal through difference phase channel

From Eq.(23), the complex differential total electric field on the Rx tag between two modulation states is

$$\Delta E_{\mathrm{rxTot}} = (\Gamma^*_{\mathrm{L},1} - \Gamma^*_{\mathrm{L},2})E_{\mathrm{sconj}} = \Delta\Gamma^*_{\mathrm{L}} E_{\mathrm{sconj}}, \tag{26}$$

which is proportional to the differential conjugate antenna reflection coefficient $\Delta\Gamma^*_{\mathrm{L}}$, the phase of which can be determined by properly choosing the antenna load impedance $Z_{\mathrm{L,i}}$ at two states. To make two different phase channel, we just need to make

$$\Delta\Gamma^{*\mathrm{ch2}}_{\mathrm{L}} = \Delta\Gamma^{*\mathrm{ch1}}_{\mathrm{L}} e^{j\theta_n}. \tag{27}$$

According to [5], to obtain the largest amplitude differences in one channel when the other has phase cancellation, the best value for θ_n is $\pi/2$. Then we can map Γ^*_{L} to Smith

Chart to get the modulation scheme needed for the two channel. To make the mapping more convenient, we will use power reflection coefficient γ_L defined by Kurokawa in [6]:

$$\gamma_{\mathrm{L,i}} = \frac{Z_{\mathrm{L,i}} - Z_A^*}{Z_{\mathrm{L,i}} + Z_A}, \mathrm{i} = 1, 2. \tag{28}$$

Then $\Gamma_{\mathrm{L}}^* = -\gamma_{\mathrm{L}}$, and $\Delta\Gamma_{\mathrm{L}}^* = -\Delta\gamma_{\mathrm{L}}$.

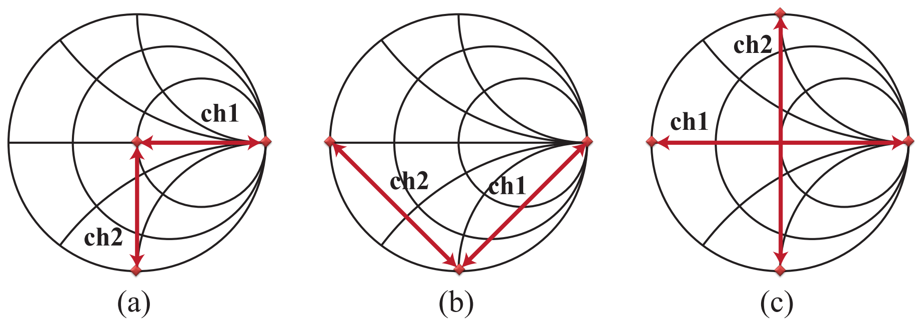

Figure 10: Three possible modulation schemes for two phase channels on normalized Smith charts in term of γ_L

Fig. 10 shows three possible modulation schemes for two phase channels on normalized Smith charts in term of γ_L. From the figure, it's very obvious that modulation scheme(c) maximize the backscatter signal strength by having the maximum amplitude of differential power reflection coefficient $\Delta\gamma_L$. Semi-passive tag or passive tag with enough storage power before backscattering can apply scheme(c) to maximize the reading distance. However, for tag implemented on CMOS, it's very hard to yield large value inductors of good quality. In this case, scheme(b) can fulfill $\pi/2$ two phase channel by only requiring RC components. Scheme(c) is good for passive tags need to harvest operating power continuously from incident field. $\Delta\gamma_L$ for three modulation schemes are shown in Table.1 as well as corresponding $\gamma_{\mathrm{L,i}}$. After locating Γ_{L}^* or γ_{L}, we can get corresponding impedance by rearranging Eq. (6) or (28):

$$Z_{\mathrm{L}} = \frac{Z_A^* - Z_A\Gamma_{\mathrm{L}}^*}{1 + \Gamma_{\mathrm{L}}^*} = \frac{Z_A^* + Z_A\gamma_{\mathrm{L}}}{1 - \gamma_{\mathrm{L}}}. \tag{29}$$

Our prototype tag implement scheme(a) for the multiphase backscattering, and the antenna impedance is $Z_A = 50/\Omega$; so corresponding $Z_{\mathrm{L,2}}^{\mathrm{ch1}}$ and $Z_{\mathrm{L,2}}^{\mathrm{ch2}}$ are ∞ (open circuit) and $-j50$ ($C \approx 3.6$ pF for 915 MHz) respectively, and $Z_{\mathrm{L,1}}^{\mathrm{ch1}} = Z_{\mathrm{L,1}}^{\mathrm{ch2}} = 50\Omega$, which is realized by matched detector.

5 Conclusion

In this report we investigated phase cancellation problem, the phenomenon of loss of baseband signal existing uniquely and ubiquitously in general BBTT communication

Table 1: Reflection coefficient for 3 modulation schemes considered

	Channel 1			Channel 2		
	$\Delta\gamma_{\mathrm{L}}$	$\gamma_{\mathrm{L},1}$	$\gamma_{\mathrm{L},2}$	$\Delta\gamma_{\mathrm{L}}$	$\gamma_{\mathrm{L},1}$	$\gamma_{\mathrm{L},2}$
scheme (a)	$1e^{j0}$	0	1	$1e^{-j\frac{\pi}{2}}$	0	-j
scheme (b)	$\sqrt{2}e^{j\frac{\pi}{4}}$	-j	1	$\sqrt{2}e^{j\frac{3\pi}{4}}$	-j	-1
scheme (c)	$2e^{j0}$	-1	1	$2e^{j\frac{\pi}{2}}$	-j	j

system. To reduce power consumption of circuit and manufacture cost, general BBTT tags utilize passive envelop detector as receiver, which can only receive the signal by detecting power incident power difference. However, there is certain possibility that the combination of CW signal and backscattered signals at different modulation states may appear the same power which is hard to be differentiated under passive detection. To evaluate this problem, instead of studying it from power perspective, we do the mathematical communication link analysis from EM field perspective. We first review the scattering theory of antenna, and derive the analytical equations for describing the EM fields in the environment under investigation and characterizing the communication link. Finally according to the properties of the field performance, we proposed a solution of multiphase backscattering, and brought up three enhanced modulation scheme based on this idea.

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