Bit and Power Loading for Optical IM/DD Transmission

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***Abstract*—Orthogonal Frequency Division Multiplexing (OFDM) is an established technique for equalization of the dispersion-dominated fiber-optical channel. The dispersive nature of the optical channel results in a frequency selective equivalent channel if double sideband (DSB) intensity modulation (IM) and direct detection (DD) is employed. In order to maintain the allpass nature of the physical channel, sideband suppression using optical filters has to be applied. We will show that by use of bit and power loading, this optical filter can be rendered unnecessary.**

I. Introduction

In the past years, Orthogonal Frequency Division Multiplexing (OFDM) has become an accepted technique for fiber-optical transmission systems due to the ease of equalization, e.g. [1]. For optical OFDM transmission with direct detection (DD), suppression of a sideband is a necessity in order to establish a successful transmission at all, regardless whether this suppression is performed in the optical domain using an optical filter or in the digital domain using a quadrature modulator structure. In this paper, we will show, that sideband suppression can be avoided if hardware complexity is shifted into the digital domain, in particular if bit and power loading is applied. Recently [2], a loading scheme for such systems was proposed, but the presented bit error performance appeared to be inferior to systems employing optical sideband suppression. We will show that systems using bit and power loading can perform nearly equal to systems using optical sideband suppression at only a small SNR penalty.

The remainder of this paper is organized as follows: Section II will introduce the system model and gives an overview on the components of the system and the effects their properties lead to. Section III will describe the premises and actual parameters of the systems under investigation. The simulation results for these systems are presented in the following section, Section V is a short excursion into the problem of iterative interference-aware loading, while the last section will conclude this paper.

II. System Model

The model of the OFDM system under consideration is depicted in Fig. 1. An analog signal $x_{\mathrm{c}}(t)$ is generated using a mostly conventional OFDM transmitter structure. The only difference is that the lower sideband subcarriers are constructed from the upper sideband subcarriers by conjugate complex extension to ensure a real valued time domain signal. This signal, scaled by drive level m, a setup parameter which can be chosen arbitrarily, modulates the intensity of an optical carrier by means of a Mach-Zehnder-Modulator (MZM) with a cosine-shaped characteristic. The operation point on this characteristic is determined by a bias u_{bias}, which also is a setup parameter. The intensity modulation of an optical carrier by a real valued baseband signal results in an optical double sideband bandpass signal. This signal passes through a physical channel described by baseband impulse response $h_{\mathrm{ch}}(t)$ incorporating the optical fiber and optional optical filters. It then is superimposed with noise $\eta(t)$ and, after being bandpass filtered, is downconverted into an electrical signal using a photo diode performing a magnitude-square operation. After discretization by an analog-digital converter, this signal is then fed into a conventional OFDM receiver structure.

The optical fiber itself has allpass characteristic (in a certain bandwidth, which, however, is significantly larger than the signal bandwidth), described in frequency domain by the baseband transfer function

$$H_{\mathrm{f}}(j\omega) = H' \cdot \mathrm{e}^{-j\tau\omega} \cdot \mathrm{e}^{jb_2\omega^2}. \tag{1}$$

The constant b_2 is denoting the cumulated dispersion, which is proportional to the physical length L of the fiber. The linear phase component is represented by τ, while phase factors and attenuation have been incorporated in the complex valued factor H'. If filtering at the detector input with $H_{\mathrm{LP}}(j\omega)$ is appended, which in baseband description is a low pass, the physical channel

$$H_{\mathrm{ch}}(j\omega) = H_{\mathrm{LP}}(j\omega) \cdot H' \cdot \mathrm{e}^{-j\tau\omega} \cdot \mathrm{e}^{jb_2\omega^2} \tag{2}$$

is obtained.

The effective channel of a direct detection system can be constructed by a real part operation of the (complex-valued) scaled physical channel impulse response $h_{\mathrm{ch}}(t)$ [3], i.e.,

$$\tilde{h}(t) = \mathrm{Re}\left\{H_0 h_{\mathrm{ch}}(t)\right\}. \tag{3}$$

H_0 is a complex-valued scaling factor which depends on the operation point and carrier phase of the system. In the following, this factor and also the factor H' can be assumed

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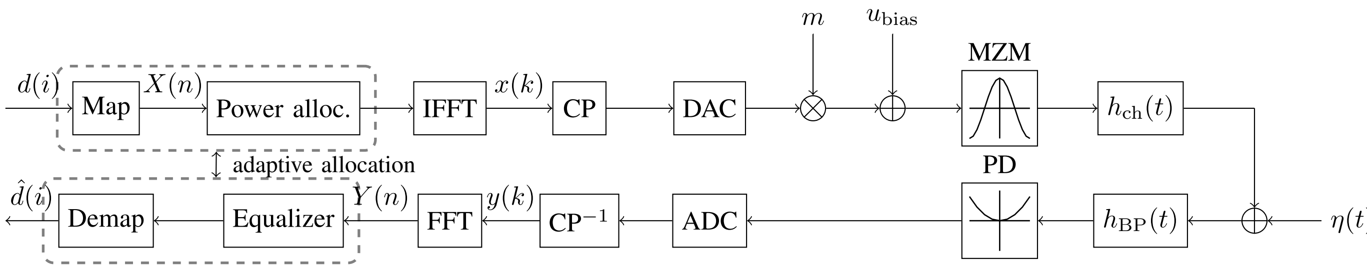

Fig. 1. Block diagram of the OFDM transmission system with bit and power loading

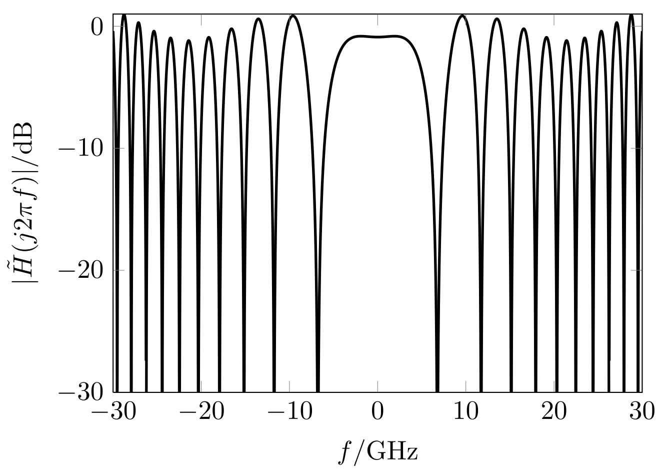

Fig. 2. Frequency response of the equivalent channel for a fiber length of 80 km and a dispersion constant of 17 ps/(nm·km)

to be equal to 1 for simplicity reasons. The schemes presented in this paper are applicable to arbitrary H_0 and H'. If $h_{\text{LP}}(t)$ is assumed to be real valued, the real part operation in time domain corresponds to a superposition with a frequency reversed complex conjugate frequency response in frequency domain, i.e., in our considered case

$$\tilde{H}(j\omega) = H_{\text{LP}}(j\omega)\mathrm{e}^{-j\tau\omega}\cos(b_2\omega^2). \tag{4}$$

Obviously, the equivalent baseband exposes a frequency selectivity, which is not present in the physical channel. An example of such a channel, as it is used in the simulations later on for a fiber length of 80 km and a dispersion constant of 17 ps/(nm·km) is depicted in Fig. 2. The ripples visible in the plot result from a non-ideal $H_{\text{LP}}(j\omega)$. The OFDM subcarriers around the channel zeros, e.g., at 6.8 or 11.7 GHz are not suitable for data transmission. In order to avoid this frequency fading and maintain the allpass characteristic of the physical bandpass channel, suppression of one sideband using an expensive optical filter is usually applied. Theoretically, perfect lower sideband suppression is achieved with a filter

$$h_{\text{SSB}}(t) = \delta_0(t) + j \cdot h_{\text{Hilbert}}(t) \tag{5}$$

where $h_{\text{Hilbert}}(t)$ is the impulse response of a Hilbert transformer and $\delta_0(t)$ is a Dirac delta function. Convolution with this filter enforces $H_{\text{ch}}(j\omega) = 0$ for $\omega < 0$, ensuring that, when superimposing of $H_{\text{ch}}(j\omega)$ with its frequency reversed complex conjugate version, the non-vanishing halves of the frequency response don't overlap. If this sideband suppression filter is inserted in the optical domain, the equivalent baseband channel is given in time domain as

$$\tilde{h}(t) = \mathrm{Re}\left\{h_{\text{SSB}}(t) * h_{\text{ch}}(t)\right\}, \tag{6}$$

which transforms into

$$\tilde{H}(j\omega) = H_{\text{LP}}(j\omega)\mathrm{e}^{-j\tau\omega}\mathrm{e}^{jb_2|\omega|\cdot\omega}. \tag{7}$$

In this case, the allpass characteristic of the physical channel is preserved.

The sideband suppression has to be performed in the optical domain, since intensity modulation always results in a double sideband bandpass signal. If sideband suppression was intended to be applied in the electrical or digital domain, a quadrature I/Q modulator would be required, which would represent a hardware expense even larger than the optical filtering intended to be rendered unnecessary by our attempts: The high frequency selectivity of the channel can be dealt with if digital signal processing is applied.

We showed that using subcarrier selection, a transmission over the frequency selective channel can be established [4], but with a SNR requirement not competitive to optical sideband suppression, making the necessity of bit and power loading obvious. The loading scheme presented in [2] for application on such frequency selective channels exposed a bit error performance that appeared to be inferior to systems employing optical sideband suppression. Independently of this work, we have investigated the applicability of a bit and power loading algorithm proposed by Krongold et al. [5] to aforementioned frequency selective optical channels.

III. Bit and Power Loading

In conventional wireline and wireless multicarrier systems, where bit and power loading algorithms are applied, the noise encountered on the subcarriers usually is entirely of external nature and not resulting from self-interference. In such cases, the loading algorithm can assume the noise power used for calculation of the "channel to noise ratio" (CNR) used as an input parameter to the optimization to be constant and independent of the subcarrier powers. In optical DD systems – unless a spectral gap reducing the spectral efficiency is used – the noise on the nth subcarrier consists of both uncorrelated zero-mean noise $W(n)$ and interference $Z(n)$ resulting from intermodulation of the subcarriers due to the nonlinearity of the overall transmission chain, whose power depends on the actual power allocation:

$$Y(n) = \tilde{H}(n)\sqrt{P(n)}X(n) + W(n) + Z(n). \qquad (8)$$

In [2], an iterative scheme is employed for calculation of the CNRs, but the actual loading algorithm that has been used for that purpose is not named and especially its convergence properties are unclear.

In this work, we will use bit and power loading as proposed by Krongold [5] for our considerations. This algorithm uses a Lagrange-multiplier bisection search to find the optimum bit and power allocation for given CNRs. Its actual mode of operation is not of interest for this paper and is therefore not reviewed here.

A. Non-linearized system setup

The Mach-Zehnder modulator used for modulation of the optical carrier uses the interferometer principle and, thus, exposes a cosine-shaped characteristic, while the photo diode at the receiver performs the aforementioned magnitude-square operation. On the overall cosine-square characteristic, an operation point has to be set up using an appropriate bias u_{bias} and modulation depth m [6]. The choice of these parameters determines the signal-to-interference ratio on the one hand and the carrier-to-sideband power ratio on the other hand, which in a power limited system both determine the overall performance significantly. The interference term $Z(n)$ in (8) consists of both interference from the cosine characteristic and optional clipping at the transmitter and inter-carrier-interference resulting from the magnitude-square operation at the receiver. While the transmitter-generated interference is analytically hard to describe and highly dependent on the operation point, the latter can exactly be constructed by self-convolution of the carrier-removed signal spectrum with its frequency-reversed complex conjugate. For any complex-valued signal $v(t)$, a magnitude-square operation can be described by

$$|v(t)|^2 = v(t) \cdot v^*(t) \;\circ\!\!-\!\!\bullet\; V(j\omega) * V^*(-j\omega). \qquad (9)$$

Given a certain power allocation, the resulting receiver-generated interference can therefore be approximated using above expression and considered as additional noise in the CNRs used for the loading algorithm. Of course this approach requires an iterative procedure whose convergence cannot be guaranteed.

For subsequent simulations, the conventional case of single sideband (SSB) transmission with optical sideband suppression shall serve as a reference. For this setup however, it is nontrivial to find the optimum operation point and parameter set. It is widely established [1] to omit the subcarriers around the carrier from allocation in order to form a spectral gap that collects the interference from (9) (often called "Offset SSB") and allocate equal powers and modulation to the remaining subcarriers. Since the subcarriers around the carrier are the least frequency selective (cf. Fig. 2), it might be an option to include these subcarriers in the loading process and accept the resulting interference. We will consider this option in our simulations.

Bias and modulation depth are parameters that have to be optimized, this can be done either analytically using worst case approximations [6] or through exhaustive search. In our case, we will use the latter approach and use the required optical signal-to-noise ratio (OSNR) for a target bit error rate (BER) of 10^{-3} as a criterion. An important aspect that has to be considered is if clipping should be applied at the transmitter side. If the signal is clipped in the extrema of the overall cosine-square characteristic, ambiguities are avoided [6], but hard clipping creates spectral components that, if passed through the physical optical channel, are reflected in interference terms that deteriorate the system performance significantly. Therefore, we will consider both cases for our simulations.

B. Linearized system setup

The nonlinearities of transmitter and/or receiver can be (partially) linearized using digital signal processing [7]. While the squared-magnitude operation at the receiver can be corrected using a square-root operation, the cosine characteristic of the Mach-Zehnder modulator can be partially linearized using an arc cosine function. Unfortunately, the transmitter linearization requires a mandatory clipping of the transmit signal, which leads to interference, as noted before. Therefore, we will consider both the case of receiver-only and full linearization for our simulations. For transmitter-side linearization, digital setup parameters m_{pre} and u_{pre} replace m and u_{bias}, these are also optimized for a target bit error rate of 10^{-3}.

IV. Simulation Results

A. Non-linearized system setup

In the simulations, two DSB OFDM systems applying bit and power loading with and without gap are compared to a single sideband OFDM system using optical sideband suppression, both with and without clipping. The setup parameters of all systems were optimized individually and are denoted in Table I. The FFT length was 2048 with twofold oversampling. Omitting DC and Nyquist frequency subcarrier, 511 of the remaining 1022 subcarriers were available for allocation, while the rest was conjugate complex extension.

The spectral gap, if used, was 255 subcarriers wide, the upper 256 subcarriers were assigned QPSK symbols in the SSB case, resulting in 512 data bits per OFDM symbol. For the loading system, the algorithm was instructed to allocate 512 bits over all 511 subcarriers in the case without gap and the upper 256 subcarriers only otherwise. The channel $h_{\mathrm{ch}}(t)$ represented 80 km of standard single mode fiber with a dispersion constant of 17 ps/(nm·km) at a carrier wavelength of 1550 nm. A cyclic prefix of length 1/8 was used, equalization was performed using an estimation based on 32

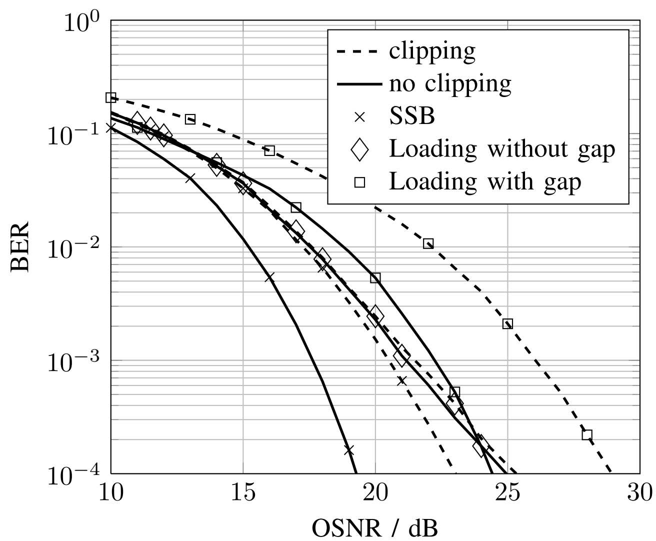

Fig. 3. Average BER with and without transmitter-side clipping for SSB and loading with and without gap using setup parameters as denoted in Table I

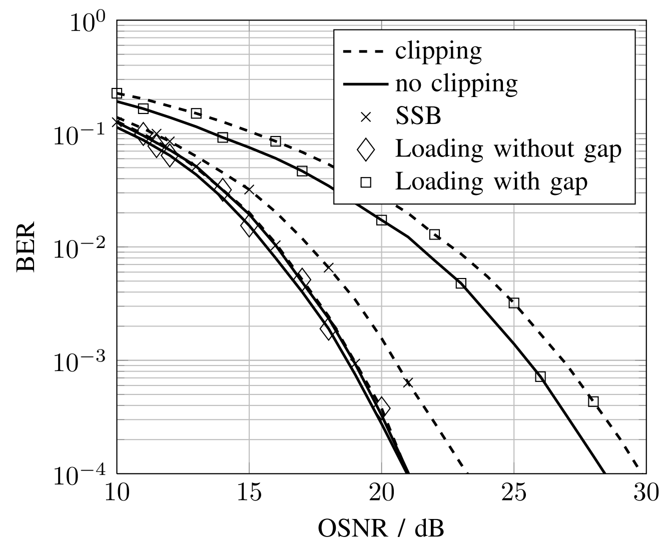

Fig. 4. Average BER with and without transmitter-side clipping for SSB and loading with and without gap using setup parameters as denoted in Table II with linearization of the receiver-side nonlinearity

TABLE I
OPTIMIZED SETUP PARAMETERS IN THE NON-LINEARIZED CASE

Scheme / Variant	u_{bias}/V_π	m/V_π
SSB with clipping	-0.9	0.2
SSB without clipping	-0.86	0.2
Loading without gap, with clipping	-0.64	0.2
Loading without gap, without clipping	-0.67	0.17
Loading with gap, with clipping	-0.7	0.17
Loading with gap, without clipping	-0.84	0.18

TABLE III
OPTIMIZED SETUP PARAMETERS IN THE CASE OF LINEARIZATION OF BOTH TRANSMITTER-SIDE AND RECEIVER-SIDE NONLINEARITIES

Scheme / Variant	u_{pre}	m_{pre}
SSB	0.08	0.1
Loading without gap	0.17	0.1
Loading with gap	0.18	0.1

different random OFDM training symbols. The average Monte Carlo simulated bit error rate curves for a bit rate of 42.8 Gb/s are depicted in Fig. 3. It can be seen that schemes applying transmitter-side clipping always perform worse than schemes without clipping. However, the difference is smallest with the loading scheme without spectral gap, whose performance is already interference-dominated, which can be seen by the reduced steepness of the curve. Nevertheless, the loading scheme without gap and clipping still performs approximately 4 dB worse at BER=10^{-3} than the optical SSB scheme.

B. Linearized system setup

The results of the simulations with linearized receiver nonlinearity are depicted in Fig. 4. Compared to the non-linearized case, it can be seen that here clipping causes a smaller SNR loss and that the case of loading without spectral gap has the same slope as the other cases. Both can be explained by the fact that the spectral self-convolution of the magnitude-square operation is avoided, reducing interference. Here, loading without clipping and without spectral gap is able to outperform the SSB scheme without clipping, but, however barely achieves the performance of SSB in the non-linearized case.

TABLE II
OPTIMIZED SETUP PARAMETERS IN THE CASE OF LINEARIZATION OF THE RECEIVER-SIDE NONLINEARITY

Scheme / Variant	u_{bias}/V_π	m/V_π
SSB with clipping	-0.88	0.2
SSB without clipping	-0.82	0.2
Loading without gap, with clipping	-0.66	0.2
Loading without gap, without clipping	-0.68	0.2
Loading with gap, with clipping	-0.65	0.18
Loading with gap, without clipping	-0.70	0.18

The simulation results with additional transmitter-side linearization are shown in Fig. 5. Compared to the case with only receiver-side linearization and clipping, an improvement can be observed for the two loading schemes, but not for the SSB scheme. However, the case of only receiver-side linearization without clipping still performs better in every case.

Wrapping up the conclusions, for the receiver-linearized case, the loading algorithm without spectral gap and clipping can compete with the SSB scheme without clipping and any linearizations at an OSNR loss of only approximately 1 dB at the target BER of 10^{-3}.

V. ITERATIVE ALLOCATION

In [2], an iterative allocation algorithm was introduced, which estimates the interference on the subcarriers resulting from the nonlinearities. However, its exact mode of operation was not detailed, especially, what nonlinear elements are considered for estimation of the interference. As mentioned before, the only interference that can be estimated precisely

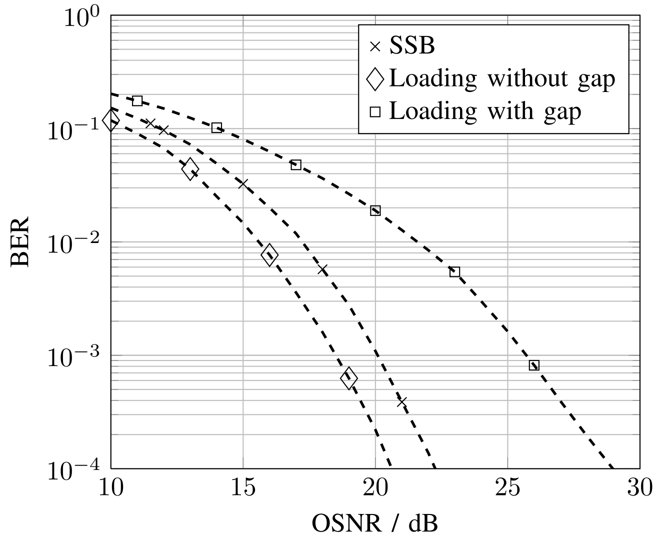

Fig. 5. Average BER with transmitter-side clipping for SSB and loading with and without gap using setup parameters as denoted in Table III with linearization of both transmitter-side and receiver-side nonlinearities

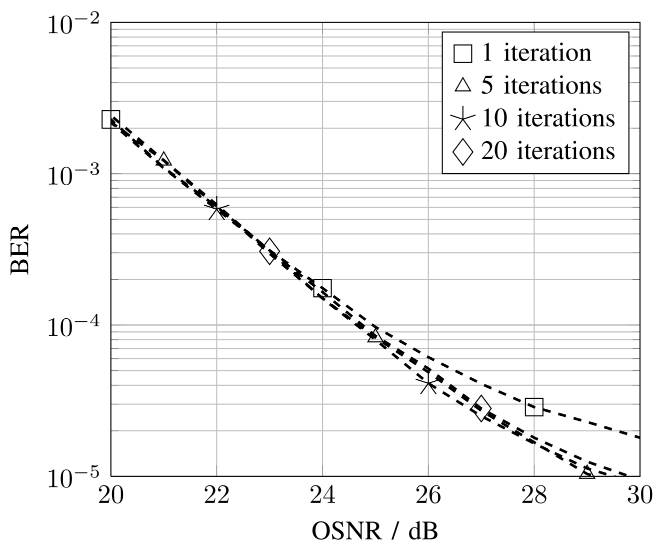

Fig. 6. Average BER loading without gap and clipping and no linearization for different number of iterations using setup parameters $u_{\mathrm{bias}}/V_\pi = -0.67$ and $m/V_\pi = 0.17$

is the interference resulting from the magnitude-square operation. Therefore, we have extended our Krongold loading scheme by an iterative concept that estimates the worst-case interference for a given bit and power allocation and regards this interference as additional noise in the calculation of the CNRs. Fig. 6 displays the resulting bit error rates for a loading scheme without spectral gap and clipping and without linearization. The setup parameters were chosen identical to the corresponding curve in Fig. 3 and Table I. It is obvious that an improvement over the non-iterative approach (denoted as "1 iteration" in the figure) is only achieved for the error-floor region and at the interesting BER region of 10^{-3}. However, the loading scheme without gap and linearization does not show a competitive performance compared to the SSB case anyhow, so improvements due to an iterative allocation would only be beneficial in the receiver-linearized case. But in that case the interference is resulting only from transmitter-side nonlinearities, whose spectral properties are hard to describe analytically and would require an exhaustive analysis.

VI. Conclusion

We have shown that, in contrast to other investigations, DSB IM/DD systems using bit and power loading can perform nearly equal to SSB IM/DD systems at the advantage of reduced hardware cost. We have also considered the case of linearized modulator and/or detector nonlinearities. Furthermore, we have shown that an iterative allocation scheme based on the resulting self-interference as proposed in the literature is not required to achieve this performance.

The loading approach presented here is not restricted to systems using single mode fibers, but to all direct detection systems with dispersive media. Especially for cost-sensitive systems employing inexpensive multi-mode fibers for metropolitan areas, the presented loading scheme might be an interesting technique.

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