MULTIRATE SIGNAL PROCESSING

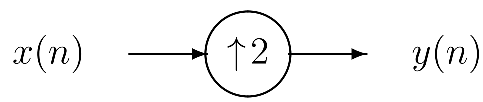
1. APPLICATIONS
2. THE UP-SAMPLER
3. THE DOWN-SAMPLER
4. RATE-CHANGING
5. INTERPOLATION
6. HALF-BAND FILTERS
7. NYQUIST FILTERS
8. THE NOBLE IDENTITIES
9. POLYPHASE DECOMPOSITION
10. EFFICIENT IMPLEMENTATION
11. POLYNOMIALS AND MULTIRATE FILTERING
12. INTERPOLATION OF POLYNOMIALS

APPLICATIONS

1. Used in A/D and D/A converters.
2. Used to change the rate of a signal. When two devices that operate at different rates are to be interconnected, it is necessary to use a rate changer between them.
3. Interpolation.
4. Some efficient implementations of single rate filters are based on multirate methods.
5. Filter banks and wavelet transforms depend on multirate methods.

THE UP-SAMPLER

The *up-sampler*, represented by the diagram,



is defined by the relation

$$y(n) = \begin{cases} x(n/2), & \text{for } n \text{ even} \\ 0, & \text{for } n \text{ odd.} \end{cases} \tag{1}$$

The usual notation is

$$y(n) = [\uparrow 2]\, x(n). \tag{2}$$

The up-sampler simply inserts zeros between samples. For example, if $x(n)$ is the sequence

$$x(n) = \{\dots,\ 3,\ \underline{5},\ 2,\ 9,\ 6,\ \dots\}$$

where the underlined number represents $x(0)$, then $y(n)$ is given by

$$y(n) = [\uparrow 2]\, x(n) = \{\dots,\ 0,\ 3,\ 0,\ \underline{5},\ 0,\ 2,\ 0,\ 9,\ 0,\ 6,\ 0,\ \dots\}.$$

Given $X(z)$, what is $Y(z)$? Using the example sequence above we directly write

$$X(z) = \cdots + 3\,z + 5 + 2\,z^{-1} + 9\,z^{-2} + 6\,z^{-3} + \cdots \tag{3}$$

and

$$Y(z) = \cdots + 3\,z^{2} + 5 + 2\,z^{-2} + 9\,z^{-4} + 6\,z^{-6} + \cdots \tag{4}$$

It is clear that

$$\boxed{Y(z) = \mathcal{Z}\{[\uparrow 2]\, x(n)\} = X(z^2).} \tag{5}$$

We can also derive this using the definition:

$$Y(z) = \sum_n y(n)\, z^{-n} \tag{6}$$

$$= \sum_{n \text{ even}} x(n/2)\, z^{-n} \tag{7}$$

$$= \sum_n x(n)\, z^{-2\,n} \tag{8}$$

$$= X(z^2). \tag{9}$$

How does up-sampling affect the Fourier transform of a signal?

The discrete-time Fourier transform of $y(n)$ is given by

$$Y(e^{j\omega}) = X(z^2)\big|_{z=e^{j\omega}} \tag{10}$$

$$= X((e^{j\omega})^2) \tag{11}$$

so we have

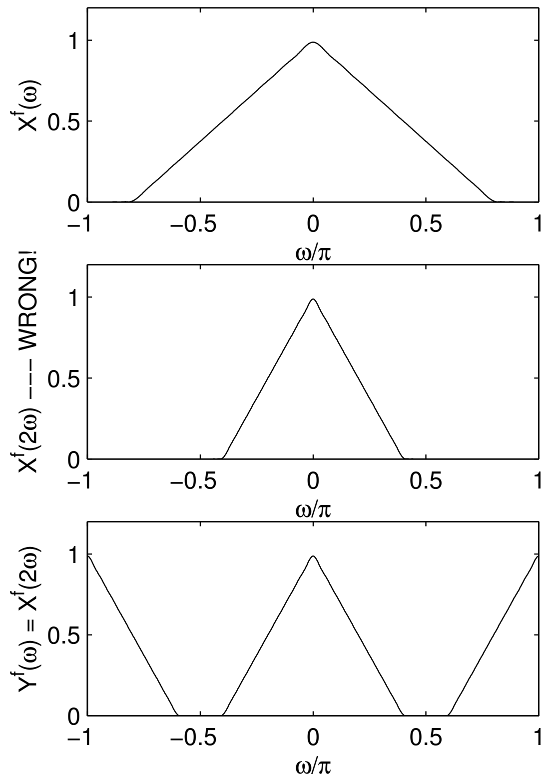
$$\boxed{Y(e^{j\omega}) = X(e^{j2\omega}).} \tag{12}$$

Or using the notation $Y^f(\omega) = Y(e^{j\omega})$, $X^f(\omega) = X(e^{j\omega})$, we have

$$\boxed{Y^f(\omega) = \text{DTFT}\{[\uparrow 2]\, x(n)\} = X^f(2\omega).} \tag{13}$$

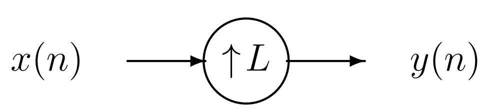
When sketching the Fourier transform of an up-sampled signal, it is easy to make a mistake. When the Fourier transform is as shown in the following figure, it is easy to incorrectly think that the Fourier transform of $y(n)$ is given by the second figure. This is not correct, because the Fourier transform is 2π-periodic. Even though it is usually graphed in the range $-\pi \le \omega \le \pi$ or $0 \le \omega \le \pi$, outside

this range it is periodic. *Because $X^f(\omega)$ is a 2π-periodic function of ω, $Y^f(\omega)$ is a π-periodic function of ω.*
The correct graph of $Y^f(\omega)$ is the third subplot in the figure.



Note that the spectrum of $X^f(\omega)$ is repeated — there is an 'extra' copy of the spectrum. This part of the spectrum is called the *spectral image*.

General case: An L-fold up-sampler, represented by the diagram,



is defined as

$$y(n) = [\uparrow L]\, x(n) = \begin{cases} x(n/L), & \text{when } n \text{ is a multiple of } L \\ 0, & \text{otherwise.} \end{cases}$$

$$(14)$$

The L-fold up-sampler simply inserts $L-1$ zeros between samples. For example, if the sequence $x(n)$

$$x(n) = \{\ldots,\ 3,\ \underline{5},\ 2,\ 9,\ 6,\ \ldots\}$$

is up-sampled by a factor $L = 4$, the result is the following sequence

$$\begin{aligned} y(n) &= [\uparrow 4]\, x(n) \\ &= \{\ldots,\ 0,\ 3,\ 0,\ 0,\ 0,\ \underline{5},\ 0,\ 0,\ 0,\ 2,\ 0,\ 0,\ 0,\ 9,\ 0,\ 0,\ 0,\ 6,\ 0,\ \ldots\}. \end{aligned}$$

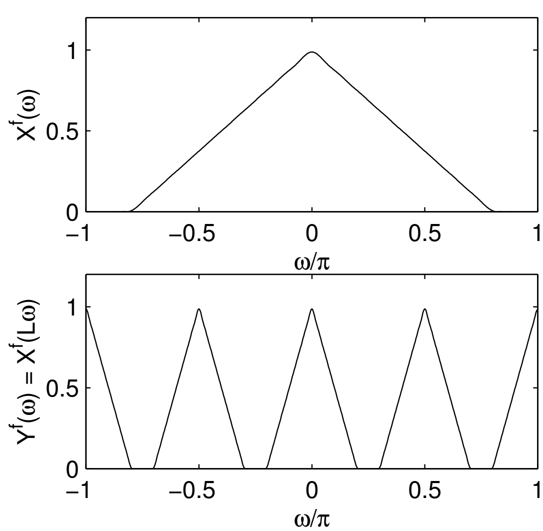
Similarly, we have

$$\boxed{Y(z) = \mathcal{Z}\{[\uparrow L]\, x(n)\} = X(z^L),} \tag{15}$$

$$\boxed{Y(e^{j\omega}) = X(e^{jL\omega}),} \tag{16}$$

$$\boxed{Y^f(\omega) = \text{DTFT}\{[\uparrow L]\, x(n)\} = X^f(L\,\omega).} \tag{17}$$

The L-fold up-sampler will create $L-1$ spectral images. For example, when a signal is up-sampled by 4, there are 3 spectral images as shown in the following figure.

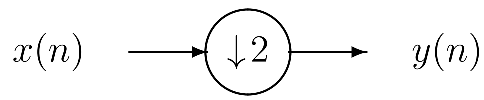


Remarks

1. No information is lost when a signal is up-sampled.
2. The up-sampler is a linear but *not* a time-invariant system.
3. The up-sampler introduces *spectral images*.

THE DOWN-SAMPLER

The down-sampler, represented by the following diagram,



is defined as

$$y(n) = x(2\,n). \tag{18}$$

The usual notation is

$$y(n) = [\downarrow 2]\, x(n). \tag{19}$$

The down-sampler simply keeps every second sample, and discards the others. For example, if $x(n)$ is the sequence

$$x(n) = \{\dots,\ 7,\ 3,\ \underline{5},\ 2,\ 9,\ 6,\ 4,\ \dots\}$$

where the underlined number represents $x(0)$, then $y(n)$ is given by

$$y(n) = [\downarrow 2]\, x(n) = \{\dots,\ 7,\ \underline{5},\ 9,\ 4,\ \dots\}.$$

Given $X(z)$, what is $Y(z)$? This is not as simple as it is for the up-sampler. Using the example sequence above we directly write

$$X(z) = \cdots + 7\,z^2 + 3\,z + 5 + 2\,z^{-1} + 9\,z^{-2} + 6\,z^{-3} + 4\,z^{-4} + \cdots \tag{20}$$

and

$$Y(z) = \cdots + 7\,z + 5 + 9\,z^{-1} + 4\,z^{-2} + \cdots \tag{21}$$

How can we express $Y(z)$ in terms of $X(z)$? Consider the sum of $X(z)$ and $X(-z)$. Note that $X(-z)$ is given by

$$X(-z) = \cdots + 7\,z^2 - 3\,z + 5 - 2\,z^{-1} + 9\,z^{-2} - 6\,z^{-3} + 4\,z^{-4} + \cdots. \tag{22}$$

The odd terms are negated. Then

$$X(z)+X(-z) = 2\cdot\left(\cdots + 7\,z^{2} + 5 + 9\,z^{-2} + 4\,z^{-4} + \cdots\right) \tag{23}$$

or

$$\boxed{X(z) + X(-z) = 2\cdot Y(z^{2})} \tag{24}$$

or

$$\boxed{Y(z) = \mathcal{Z}\left\{[\downarrow 2]\, x(n)\right\} = \frac{X(z^{\frac{1}{2}}) + X(-z^{\frac{1}{2}})}{2}} \tag{25}$$

How does down-sampling affect the Fourier transform of a signal?

The discrete-time Fourier transform of $y(n)$ is given by

$$Y(e^{j\omega}) = \left.\frac{X(z^{\frac{1}{2}}) + X(-z^{\frac{1}{2}})}{2}\right|_{z=e^{j\omega}} \tag{26}$$

$$= \frac{1}{2}\cdot\left(X(e^{j\frac{\omega}{2}}) + X(-e^{j\frac{\omega}{2}})\right) \tag{27}$$

$$= \frac{1}{2}\cdot\left(X(e^{j\frac{\omega}{2}}) + X(e^{-j\pi}\, e^{j\frac{\omega}{2}})\right) \tag{28}$$

$$= \frac{1}{2}\cdot\left(X(e^{j\frac{\omega}{2}}) + X(e^{j(\frac{\omega}{2}-\pi)})\right) \tag{29}$$

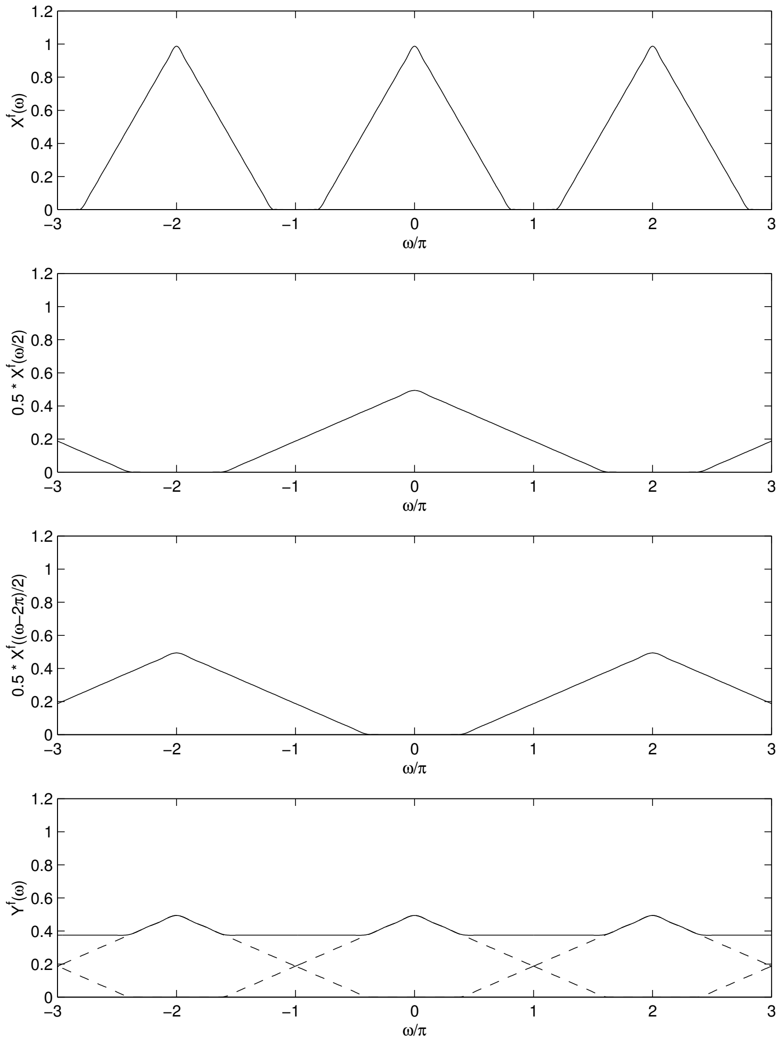
$$= \frac{1}{2}\cdot\left(X^{f}(\frac{\omega}{2}) + X^{f}(\frac{\omega-2\pi}{2})\right) \tag{30}$$

$$\boxed{Y^{f}(\omega) = \text{DTFT}\left\{[\downarrow 2]\, x(n)\right\} = \frac{1}{2}\cdot\left(X^{f}(\frac{\omega}{2}) + X^{f}(\frac{\omega-2\pi}{2})\right)} \tag{31}$$

where we have used the notation $Y^{f}(\omega) = Y(e^{j\omega})$, $X^{f}(\omega) = X(e^{j\omega})$.

Note that because $X^{f}(\omega)$ is periodic with a period of 2π, the functions $X^{f}(\frac{\omega}{2})$ and $X^{f}(\frac{\omega-2\pi}{2})$ are each periodic with a period of 4π.

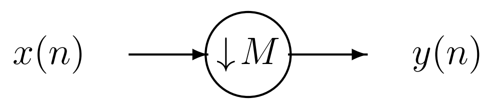
But as $Y^f(\omega)$ is the Fourier transform of a signal, it must be 2π-periodic. What does $Y^f(\omega)$ look like? It is best illustrated with an example.



Notice that while the two terms $X^f(\frac{\omega}{2})$ and $X^f(\frac{\omega-2\pi}{2})$ are 4π-periodic, because one is shifted by 2π, their sum is 2π-periodic, as a Fourier transform must be.

Notice that when a signal $x(n)$ is down-sampled, the spectrum $X^f(\omega)$ may overlap with adjacent copies, depending on the specific shape of $X^f(\omega)$. This overlapping is called *aliasing*. When aliasing occurs, the signal $x(n)$ can not in general be recovered after it is down-sampled. In this case, information is lost by the down-sampling. If the spectrum $X^f(\omega)$ were zero for $\pi/2 \leq |\omega| \leq \pi$, then no overlapping would occur, and it would be possible to recover $x(n)$ after it is down-sampled.

General case: An M-fold down-sampler, represented by the diagram,



is defined as

$$y(n) = x(M\,n). \tag{32}$$

The M-fold down-sampler keeps only every M^{th} sample. For example, if the sequence $x(n)$

$$x(n) = \{\ldots,\ 8,\ 7,\ 3,\ \underline{5},\ 2,\ 9,\ 6,\ 4,\ 2,\ 1,\ \ldots\}$$

is down-sampled by a factor $M = 3$, the result is the following sequence

$$y(n) = [\downarrow 3]\, x(n) = \{\ldots,\ 8,\ \underline{5},\ 6,\ 1,\ \ldots\}.$$

Similarly, we have

$$\boxed{Y(z) = \mathcal{Z}\left\{[\downarrow M]\, x(n)\right\} = \frac{1}{M}\sum_{k=0}^{M-1} X(W^k z^{\frac{1}{M}})} \tag{33}$$

where

$$W = e^{j\frac{2\pi}{M}}, \tag{34}$$

and

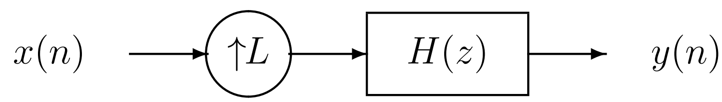
$$\boxed{Y^f(\omega) = \mathsf{DTFT}\left\{[\downarrow M]\, x(n)\right\} = \frac{1}{M} \sum_{k=0}^{M-1} X^f\left(\frac{\omega - 2\pi\, k}{M}\right).} \quad (35)$$

Remarks

1. In general, information is lost when a signal is down-sampled.
2. The down-sampler is a linear but *not* a time-invariant system.
3. In general, the down-sampler causes *aliasing*.

The up-sampler and down-sampler are usually used in combination with filters, not by themselves. For example, to change the rate of a signal, it is necessary to employ low-pass filters in addition to the up-sampler and down-sampler.

The following system is used for interpolation.

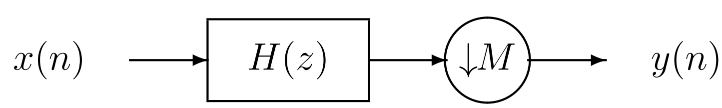


The combined up-sampling and filtering can be written as

$$y(n) = ([\uparrow L]\, x(n)) * h(n) = \sum_k x(k)\, h(n - L\,k). \tag{36}$$

The filter fills in the zeros that are introduced by the up-sampler. Equivalently, it is designed to remove the *spectral images*. It should be a low-pass filter with a cut-off frequency $\omega_o = \pi/L$. In this context, the low-pass filter is often called an *interpolation filter*.

The following system is used for decimation.

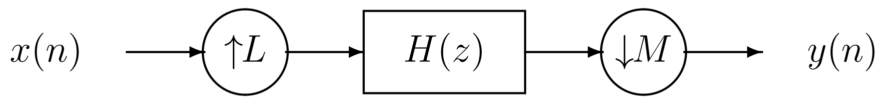


The combined filtering and down-sampling can be written as

$$y(n) = [\downarrow M]\, (x(n) * h(n)) = \sum_k x(k)\, h(M\,n - k). \tag{37}$$

The filter is designed to avoid *aliasing*. It should be a low-pass filter with a cut-off frequency $\omega_o = \pi/M$. In this context, the low-pass filter is often called an *anti-aliasing filter*.

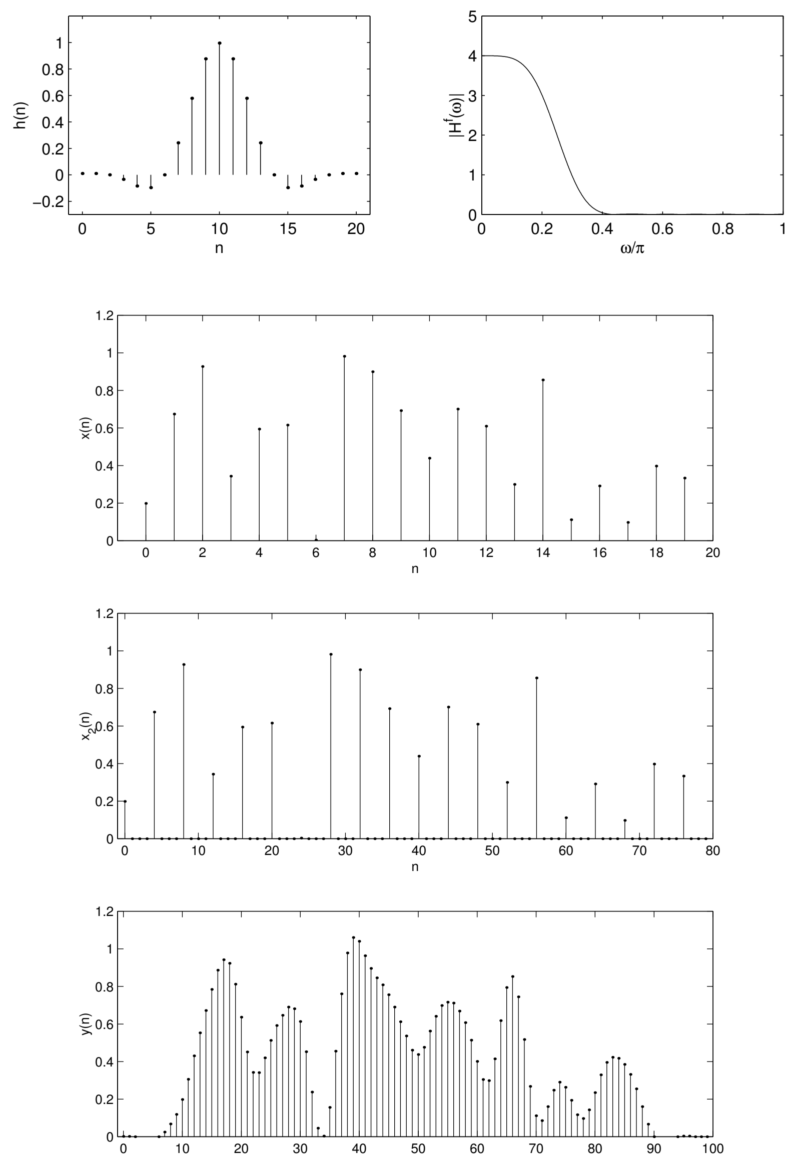
A rate changer for a fractional change (like 2/3) can be obtained by cascading an interpolation system with a decimation system. Then, instead of implementing two separate filters in cascade, one can implement a single filter. Structure for rational rate changer:



The filter is designed to both eliminate spectral images and to avoid aliasing. The cascade of two ideal low-pass filters is again a low-pass filter with a cut-off frequency that is the minimum of the two cut-off frequencies. So, in this case, the cut-off frequency should be

$$\boxed{\omega_o = \min\left\{\frac{\pi}{L}, \frac{\pi}{M}\right\}.} \tag{38}$$

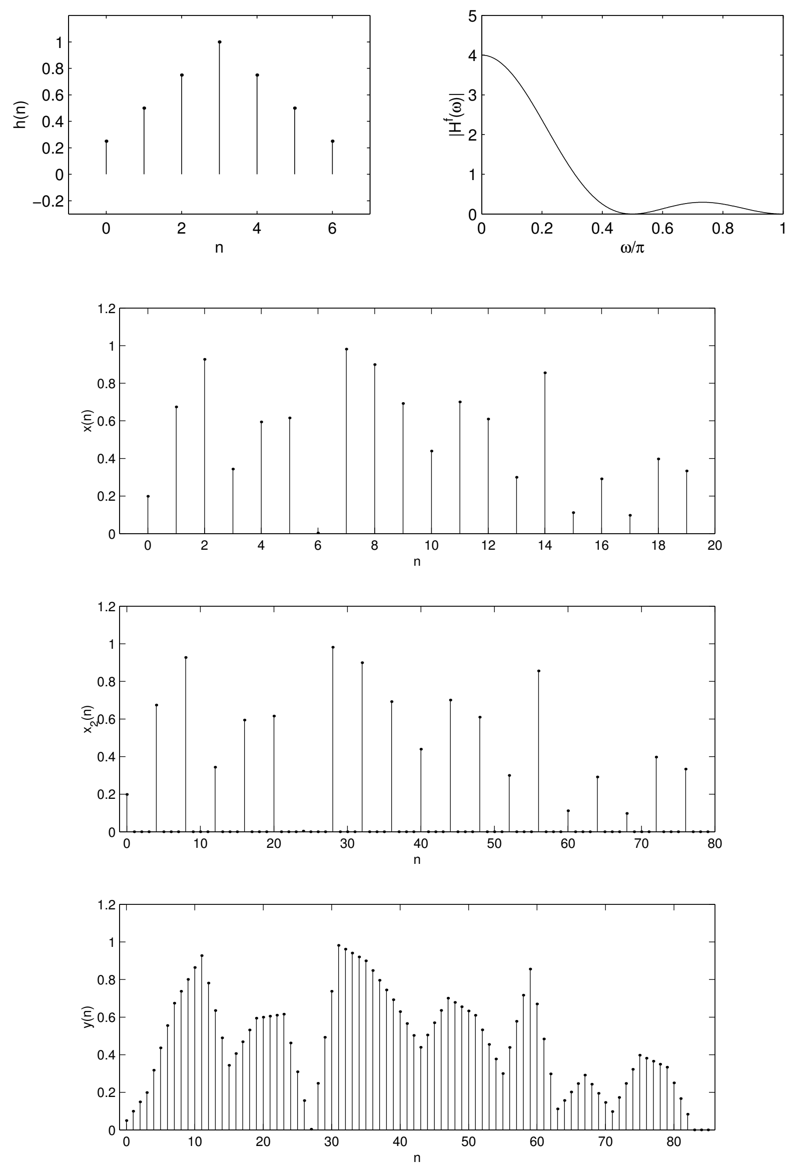
In this example, we interpolate a signal $x(n)$ by a factor of 4, using the interpolation system described above. We use a linear-phase Type I FIR lowpass filter of length 21 to follow the 4-fold up-sampler. Note that because the filter is causal, a delay in introduced by the interpolation system. $y(n)$ could be aligned with $x(n)$ by shifting it.



This time we use a filter of length 7,

$$h(n) = \frac{1}{4} \cdot \{1,\ 2,\ 3,\ 4,\ 3,\ 2,\ 1\}. \tag{39}$$

Note that this filter has the effect of implementing *linear* interpolation between the existing samples $x(n)$. The result is rather poor — the signal $y(n)$ is not very smooth. Similarly, *quadratic* interpolation can be implemented by using an appropriate filter $h(n)$.



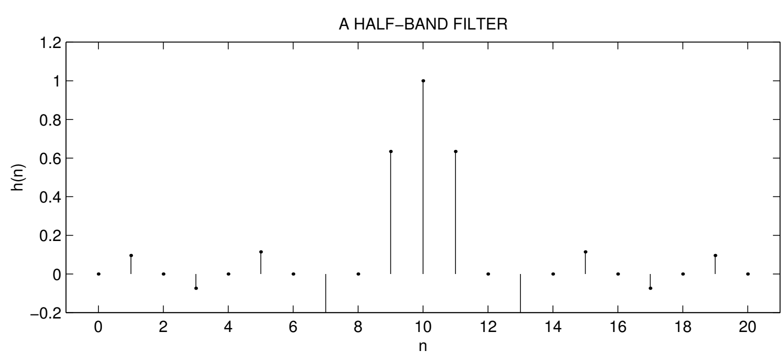
When interpolating a signal $x(n)$, the interpolation filter $h(n)$ will in general change the samples of $x(n)$ in addition to filling in the zeros. It is natural to ask if the interpolation filter can be designed so as to preserve the original samples $x(n)$.
To be precise, if

$$y(n) = h(n) * [\uparrow 2]\, x(n)$$

then can we design $h(n)$ so that $y(2n) = x(n)$, or more generally, so that $\boxed{y(2n + n_o) = x(n)}$?
It turns out that this is possible. When interpolating by a factor of 2, if $h(n)$ is a *half-band*, then it will not change the samples $x(n)$.
A n_o-centered half-band filter $h(n)$ is a filter that satisfies

$$h(n) = \begin{cases} 1, & \text{for } n = n_o \\ 0, & \text{for } n = n_o \pm 2, 4, 6, \ldots \end{cases} \tag{40}$$

That means, every second value of $h(n)$ is zero, except for one such value, as shown in the figure.



In the figure, the center point is $n_o = 10$. The definition of a half-band filter can be written more compactly using the Kronecker delta function $\delta(n)$. **Half-band filter:**

$$\boxed{h(2\,n + n_o) = \delta(n),} \tag{41}$$

when $n_o = 0$, we get simply:

$$\boxed{h(2\,n) = \delta(n).} \tag{42}$$

Note that the transfer function of a half-band filter (centered at $n_o = 0$) can be written as

$$H(z) = 1 + z^{-1}\,H_1(z^2). \tag{43}$$

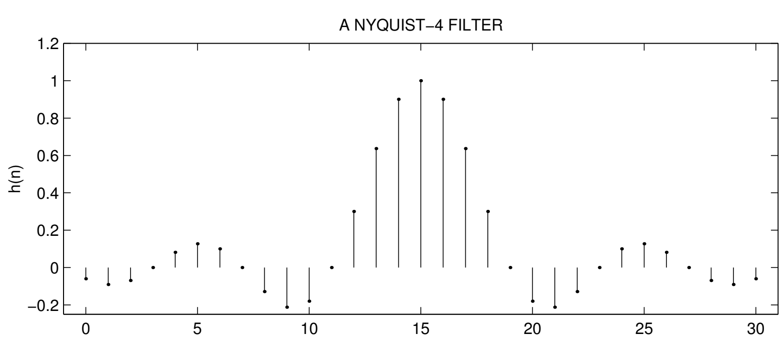
Here $H_1(z)$ contains the odd samples of $h(n)$.

NYQUIST FILTERS

When interpolating a signal $x(n)$ by a factor L, the original samples of $x(n)$ are preserved if the interpolation filter $h(n)$ is a *Nyquist-L* filter. A Nyquist-L filter simply generalizes the notion of the half-band filter to $L > 2$. A (0-centered) Nyquist-L filter $h(n)$ is one for which

$$\boxed{h(L\,n) = \delta(n).} \tag{44}$$

A Nyquist-4 filter is shown in the following figure.



THE NOBLE IDENTITY FOR THE UP-SAMPLER

The two following equivalences are sometimes called the *noble* identities.

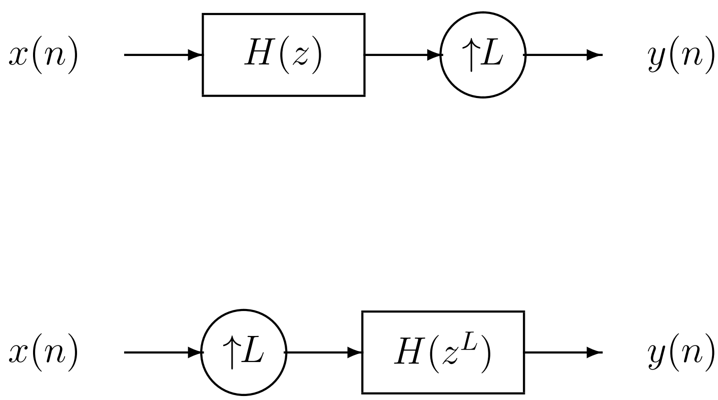
Can you reverse the order of an up-sampler and a filter?

Yes and no — it depends. There are two cases.

1. If the up-sampler comes *after* the filter, then you can reverse the order of the filter and the up-sampler, but the filter needs to be modified as shown in the figure.

2. If the up-sampler comes *before* the filter, then you can *not* reverse their order unless the filter is of the special form $H(z^L)$.

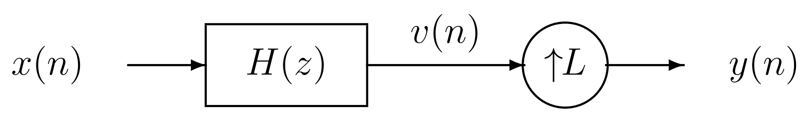
This can be summarized by the following figure.



Equivalently:

$$[\uparrow L]\ (h(n) * x(n)) = [\uparrow L]\, h(n) * [\uparrow L]\, x(n)$$

This identity is most easily derived using the Z-transform and equation (15). In the following figure the intermediate signal $v(n)$ is shown.



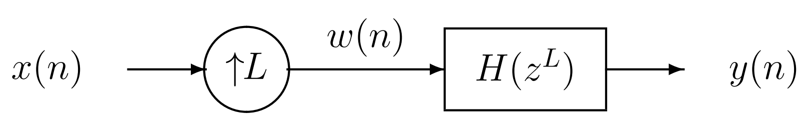
Then, using the Z-transform, we have

$$V(z) = H(z)\,X(z) \qquad \text{and} \qquad Y(z) = V(z^L)$$

and therefore,

$$Y(z) = H(z^L)\,X(z^L).$$

Now consider the system that we claim to be equivalent. In the following figure the intermediate signal $w(n)$ is shown.



Then, using the Z-transform, we have

$$W(z) = X(z^L) \qquad \text{and} \qquad Y(z) = H(z^L)\,W(z)$$

and therefore,

$$Y(z) = H(z^L)\,X(z^L).$$

This shows that the systems are equivalent.

THE NOBLE IDENTITY FOR THE DOWN-SAMPLER

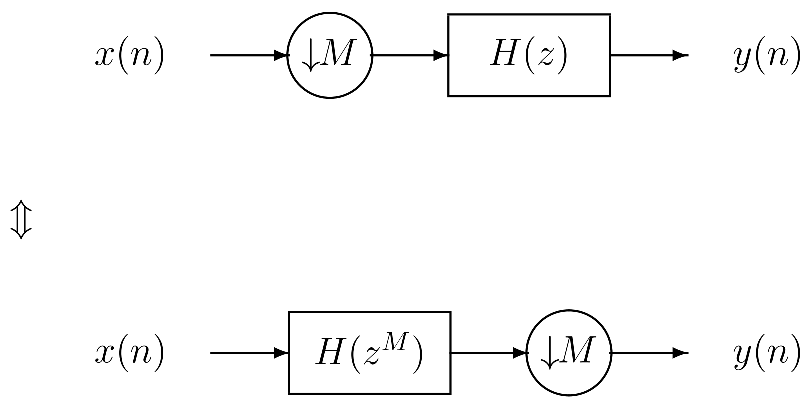
Can you reverse the order of an down-sampler and a filter?

Yes and no — it depends. There are two cases.

1. If the down-sampler comes *before* the filter, then you can reverse the order of the filter and the down-sampler, but the filter needs to be modified as shown in the figure.

2. If the down-sampler comes *after* the filter, then you can *not* reverse their order unless the filter is of the special form $H(z^M)$.

This can be summarized by the following figure.

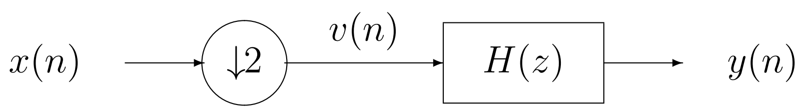


Equivalently:

$$h(n) * [\downarrow M]\, x(n) = [\downarrow M]\ ([\uparrow M]\, h(n) * x(n))$$

THE NOBLE IDENTITY FOR THE DOWN-SAMPLER

For convenience, we prove it just for $M = 2$. This identity is most easily derived using the Z-transform and equation (25). In the following figure the intermediate signal $v(n)$ is shown.



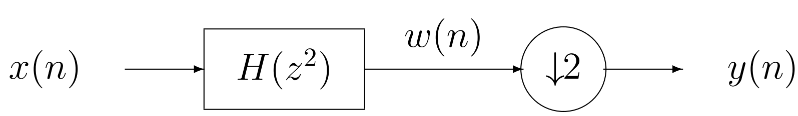
Then, using the Z-transform, we have

$$V(z) = \frac{1}{2} X(z^{\frac{1}{2}}) + \frac{1}{2} X(-z^{\frac{1}{2}}) \qquad \text{and} \qquad Y(z) = H(z)\, V(z)$$

and therefore,

$$Y(z) = \frac{1}{2} H(z)\, X(z^{\frac{1}{2}}) + \frac{1}{2} H(z)\, X(-z^{\frac{1}{2}})$$

Now consider the system that we claim to be equivalent. In the following figure the intermediate signal $w(n)$ is shown.



Then, using the Z-transform, we have

$$W(z) = H(z^2)\, X(z) \qquad \text{and} \qquad Y(z) = \frac{1}{2} W(z^{\frac{1}{2}}) + \frac{1}{2} W(-z^{\frac{1}{2}})$$

and therefore,

$$Y(z) = \frac{1}{2} H(z)\, X(z^{\frac{1}{2}}) + \frac{1}{2} H(z)\, X(-z^{\frac{1}{2}}).$$

This shows that the systems are equivalent.

The polyphase decomposition of a signal is simply the even and odd samples,

$$x_0(n) = x(2\,n) \tag{45}$$

$$x_1(n) = x(2\,n+1). \tag{46}$$

Then the Z-transform $X(z)$ is given by

$$\boxed{X(z) = X_0(z^2) + z^{-1}\,X_1(z^2)} \tag{47}$$

where $X_0(z)$ and $X_1(z)$ are the Z-transforms of $x_0(n)$ and $x_1(n)$. For example, if $x(n)$ is:

$$x(n) = \{\underline{3},\ 1,\ 5,\ 6,\ 2,\ 4,\ -3,\ 7\}$$

then the polyphase components are

$$x_0(n) = \{\underline{3},\ 5,\ 2,\ -3\} \tag{48}$$

$$x_1(n) = \{\underline{1},\ 6,\ 4,\ 7\}. \tag{49}$$

The Z-transforms for this example are given by

$$X(z) = 3 + z^{-1} + 5\,z^{-2} + 6\,z^{-3} + 2\,z^{-4} + 4\,z^{-5} - 3\,z^{-6} + 7\,z^{-7}$$

$$X_0(z) = 3 + 5\,z^{-1} + 2\,z^{-2} - 3\,z^{-3}$$

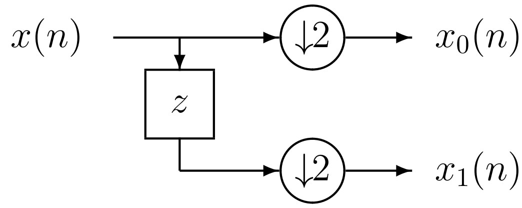
$$X_1(z) = 1 + 6\,z^{-1} + 4\,z^{-2} + 7\,z^{-3}.$$

In general, $X_0(z)$ and $X_1(z)$ can be obtained from $X(z)$ as,

$$X_0(z^2) = \frac{1}{2}\,(X(z) + X(-z)) \tag{50}$$

$$X_1(z^2) = \frac{z}{2}\,(X(z) - X(-z))\,. \tag{51}$$

The polyphase components $x_0(n)$, $x_1(n)$ can be obtained with the following structure.



General case: An M-component polyphase decomposition of $x(n)$ is given by

$$x_0(n) = x(M\,n) \tag{52}$$

$$x_1(n) = x(M\,n+1) \tag{53}$$

$$\vdots \tag{54}$$

$$x_{M-1}(n) = x(M\,n+M-1). \tag{55}$$

The Z-transform $X(z)$ is then given by

$$\boxed{X(z) = X_0(z^M) + z^{-1}\,X_1(z^M) + \cdots + z^{-(M-1)}\,X_{M-1}(z^M)} \tag{56}$$

where $X_i(z)$ is the Z-transform of $x_i(n)$. The polyphase component $X_i(z)$ can be found from $X(z)$ with

$$X_i(z^M) = \frac{z^i}{M} \sum_{k=0}^{M-1} W^{ik}\,X(W^k\,z) \tag{57}$$

where

$$W = e^{j\frac{2\pi}{M}}. \tag{58}$$

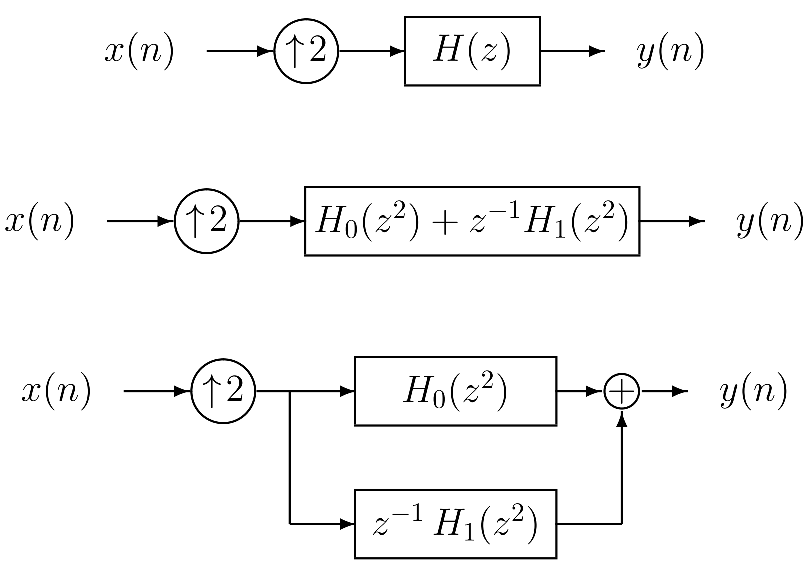
EFFICIENT IMPLEMENTATION

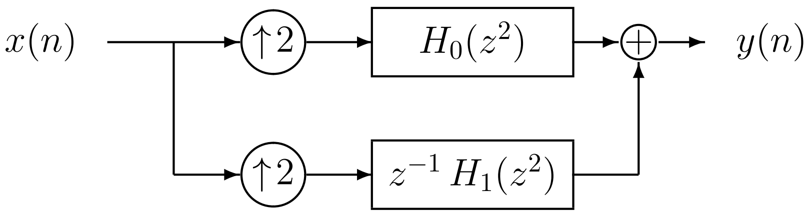
The noble identities and the polyphase decomposition can be used together to obtain efficient structures. Consider again the system for interpolation: an up-sampler is followed by a filter. In this system, the up-sampler inserts zeros between the samples $x(n)$. There are two disadvantages.

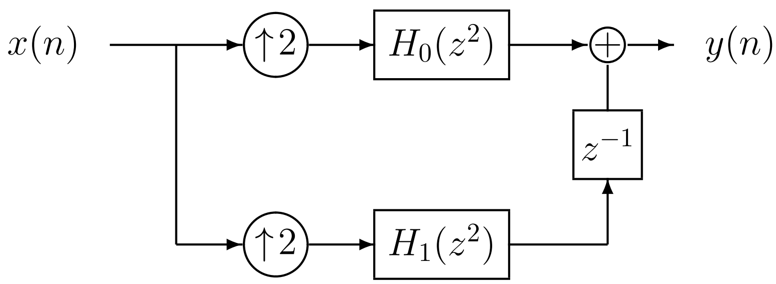
1. Half the samples of the input to the filter are zero. That means the filter is doing unnecessary computations (multiplications by zero, adding zeros).

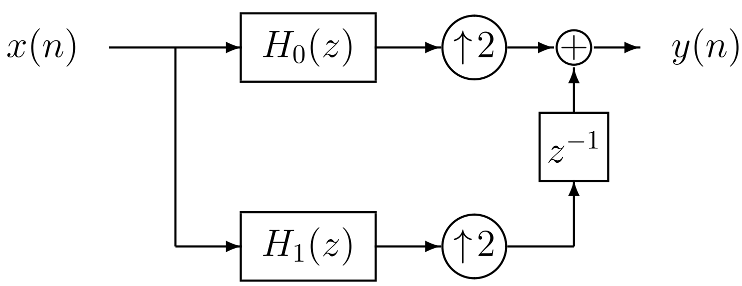
2. The filter operates at the higher rate.

A more efficient implementation can be obtained by writing the filter in polyphase form, and then using the noble identities. This is done through the following transformation of the block diagram.





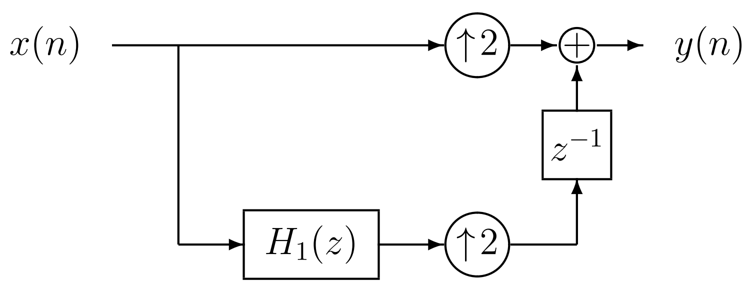




Note that in the last block diagram, the filters operate at the slower rate, and the filter inputs are not zero. Also note that the filters $h_0(n)$, $h_1(n)$ are each half the length of the original filter $h(n)$. The adding node in the last diagram does not incur any actual additions — it implements an interleaving of the two branches.

HALF-BAND CASE

If $h(n)$ is a half-band filter, then the polyphase component $H_0(z)$ is 1 (assuming the half-band filter is centered at $n_o = 0$). In this case, the block diagram becomes more simple as shown.



A (discrete-time) polynomial signal $x(n)$ is a signal of the form

$$x(n) = c_0 + c_1\, n + c_2\, n^2 + \cdots + c_d\, n^d.$$

The degree is d. The set of polynomial signals of degree d or less is denoted by $\mathcal{P}_d$.

Consider a system described by the rule

$$y(n) = x(n) - x(n-1).$$

This system gives the *first difference* of the signal $x(n)$. It has the impulse response

$$h(n) = \delta(n) - \delta(n-1),$$

and the transfer function

$$H(z) = 1 - z^{-1}$$

and so we can write

$$y(n) = h(n) * x(n)$$

or

$$Y(z) = \left(1 - z^{-1}\right)\, X(z).$$

Clearly if $x(n)$ is a constant signal ($x(n) = c$, so we can write $x(n) \in \mathcal{P}_0$), then the first difference of $x(n)$ is identically zero,

$$Y(z) = \left(1 - z^{-1}\right)\, X(z) = 0 \quad \text{for} \quad x(n) \in \mathcal{P}_0.$$

Moreover, the first difference $Y(z)$ is identically zero *only* if $x(n)$ is a constant signal.

Similarly, if $x(n)$ is a ramp signal ($x(n) = c_0 + c_1\, n$, so we can write $x(n) \in \mathcal{P}_1$), then the first difference is a constant signal. Therefore the *second difference*, (defined as the first difference of the first difference), must be identically zero. Writing this using the Z-transform gives

$$\left(1 - z^{-1}\right)^2 X(z) = 0 \quad \text{for} \quad x(n) \in \mathcal{P}_1.$$

Moreover, the second difference of $x(n)$ is identically zero *only* if $x(n)$ is of the form $c_0 + c_1\, n$. Therefore, the set of first degree polynomial signals $\mathcal{P}_1$ is exactly the set of signals that is annihilated by $(1 - z^{-1})^2$.

Similarly, if $x(n)$ is a polynomial signal of degree d, then

$$Y(z) = \left(1 - z^{-1}\right)^{d+1} X(z) = 0 \quad \text{for} \quad x(n) \in \mathcal{P}_d.$$

or equivalently,

$$y(n) = \underbrace{h(n) * h(n) * \cdots * h(n)}_{d+1 \text{ terms}} * x(n) = 0 \quad \text{for} \quad x(n) \in \mathcal{P}_d.$$

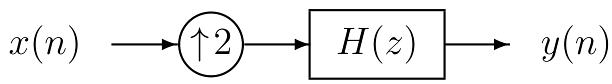
Moreover, $y(n) = 0$ *only* if $x(n)$ has the form $x(n) = c_0 + c_1\, n + c_2\, n^2 + \cdots + c_d\, n^d$.

Therefore we have the following result.

$$x(n) \in \mathcal{P}_d \quad \Longleftrightarrow \quad \left(1 - z^{-1}\right)^{d+1} X(z) = 0$$

INTERPOLATION OF POLYNOMIALS

We saw before that the interpolation of discrete-time signals can be carried out by using an upsampler together with a filter. For interpolation by a factor of two (2X interpolation) we have the following diagram.



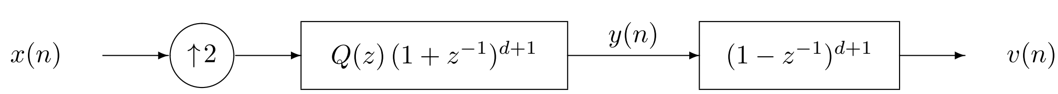
Suppose $x(n)$ is a polynomial signal of degree d. Then it is natural to ask that $y(n)$ also be a polynomial signal of degree d. But for just any filter $h(n)$ that will not be the case. What condition must $h(n)$ satisfy, to ensure that $y(n)$ is also a polynomial signal of degree d?

It turns out that if $(1+z^{-1})^{d+1}$ is a factor of $H(z)$,

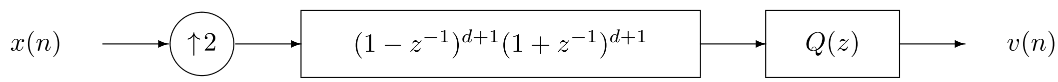
$$H(z) = Q(z)\,(1+z^{-1})^{d+1}$$

then $y(n) \in \mathcal{P}_d$ whenever $x(n) \in \mathcal{P}_d$. This can be verified using the boxed result on the previous page together with the noble identity, as we will now show.

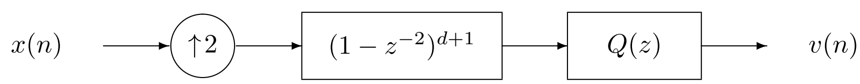
We may determine if $y(n) \in \mathcal{P}_d$ by filtering $y(n)$ with the transfer function $(1-z^{-1})^{d+1}$ and checking that the result is zero. If the signal $v(n)$ in the following figure is zero, then we know that $y(n) \in \mathcal{P}_d$ as explained earlier.



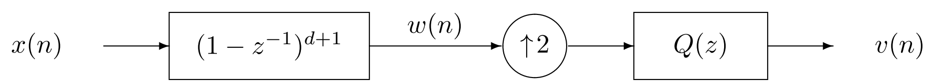
We can rearrange the order of the filters to get the following diagram which is an equivalent structure (end-to-end).



Recognizing that $(1-z^{-1})\,(1+z^{-1}) = 1-z^{-2}$ we get the following diagram.



Observing that the transfer function $(1-z^{-2})^{d+1}$ is a function of z^2, and using the noble identity for upsampling, we get the final diagram.



As explained above, if $x(n) \in \mathcal{P}_d$, then $w(n) = 0$ and therefore $v(n) = 0$. Because $v(n) = 0$, we know that $y(n) \in \mathcal{P}_d$.
In other words, if $H(z) = Q(z)\,(1+z^{-1})^{d+1}$ then when it is used for the 2X interpolation, it preserves $\mathcal{P}_d$, the set of polynomial signals of degree d.

$$H(z) = Q(z)\,(1+z^{-1})^{d+1} \iff h(n) * [\uparrow 2]\, x(n) \text{ preserves } \mathcal{P}_d$$

It should be said that in the interpolation structure above, even if $H(z)$ is chosen so that $y(n)$ is ensured to be a polynomial signal of degree d like $x(n)$ is, it does not mean that $y(2n) = x(n)$. That is only true when the filter $H(z)$ is in addition a half-band filter, as discussed above.

How should the condition above be modified if we are interpolating by a factor L rather than just by a factor of 2? If we guess that $H(z)$ should be of the form $H(z) = Q(z)\,R(z)$ and follow the same procedure used above, we will see that we will want the product $R(z)\,(1-z^{-1})^{d+1}$ to be equal to $(1-z^{-L})^{d+1}$. For in that case, we could again exchange the order of the (L-fold) upsampler and this term. This gives

$$R(z)\,(1-z^{-1})^{d+1} = (1-z^{-L})^{d+1}$$

or

$$R(z) = \frac{(1-z^{-L})^{d+1}}{(1-z^{-1})^{d+1}} = \left[1+z^{-1}+z^{-2}+\cdots+z^{-(L-1)}\right]^{d+1}$$

where we have used

$$(1-z)\,(1+z+z^2+\cdots+z^{L-1}) = 1-z^L.$$

$$H(z) = Q(z)\,\left[1+z^{-1}+z^{-2}+\cdots+z^{-(L-1)}\right]^{d+1}$$

$$\iff \qquad h(n) * [\uparrow L]\,x(n) \;\; \text{preserves} \;\; \mathcal{P}_d$$

It should be said that in the LX interpolation structure, even if $H(z)$ is chosen so that $y(n)$ is ensured to be a polynomial signal of degree d like $x(n)$ is, it does not mean that $y(Ln) = x(n)$. That is only true when the filter $H(z)$ is in addition a Nyquist-L filter.