Algorithmic Scheme-Integrated Bandwidth Compensatory Reconstruction Filter of Digital Signal Processing System

Rodjarin Boontawan\textsuperscript{1}, Siraphop Tooprakai\textsuperscript{1}, Kobchai Dejhan\textsuperscript{1} and Surapun Yimman\textsuperscript{2}

\textsuperscript{1}Faculty of Engineering, King Mongkut’s Institute of Technology Ladkrabang, Bangkok 10520, Thailand
\textsuperscript{2}Faculty of Applied Science, King Mongkut’s University of Technology North Bangkok 10800, Thailand

E-mail: rodjarinb@yahoo.com, siraphop@telecom.kmitl.ac.th, kobchai@telecom.kmitl.ac.th, sym4412@gmail.com

Abstract This research has proposed a novel algorithmic scheme of infinite impulse response (IIR) filter for bandwidth compensatory in digital signal processing (DSP) system based on the pole-zero placement technique. The scheme development started with the determination of the analog output or reconstruction filter gains at the zero, cut-off and Nyquist frequencies and then their respective nonlinear system equations. The Newton-Raphson method was subsequently applied for the positions of pole and zero and the gain of the output filter. The constants were re-substituted in the initial transfer function and an inversion carried out. To verify the applicability of the proposed scheme, simulations were carried out at two cut-off frequencies (i.e. 1.5 and 2kHz) and three sampling frequencies (i.e. 10, 12 and 14kHz). To further validate, experiments were undertaken whereby the algorithmic scheme was applied to a digital signal processing system. The simulation and experimental results revealed that the scheme implementation could effectively address the attenuation phenomenon, compensate the system bandwidth as well as enhance the time response of the output signal.

Keywords: bandwidth compensation, reconstruction filter, Newton-Raphson method, pole-zero placement

1. Introduction

A typical digital signal processing (DSP) system consists of an input filter, an analog-to-digital converter (A/D), a digital processor, a digital-to-analog converter (D/A) and an output or reconstruction filter \cite{1-4}. In operation, the DSP system bandwidth is governed by the D/A sampling property and the output filter. Most commonly available D/A converters are of zero order hold type whose magnitude response is of sinc envelop function. To tackle the sinc envelope issue requires the application of either the pre-equalizing or post-equalizing technique \cite{5,6}. In the pre-equalizing technique, a finite impulse response filter (FIR) with high order is deployed before the D/A converter, while the post-equalizing technique enquires an analog active filter subsequent to the reconstruction filter \cite{7}.

In \cite{8}, the authors proposed a bandwidth compensation scheme based on the analog transfer function of the output filter whereby the function was first converted into a discrete time system using the approximation of derivative technique and subsequently inverted for compensation of the DSP system bandwidth. However, the method suffers from low accuracy due to the utilization of the approximation of derivative technique and the subsequent discrepancy between the magnitude responses of the discrete-time system and the analog output filter. The discrepancy in turn contributes to the unsatisfactory bandwidth compensation of the DSP system.

To address the magnitude response discrepancy, this research has thus proposed a novel scheme of IIR filter based on the pole-zero placement technique. In the scheme development, the gains of the output filter at the zero, cut-off and Nyquist frequencies were determined and subsequently substituted in the initial transfer function for their respective nonlinear system equations prior to the application of the Newton-Raphson method for the positions of pole (\(p_i\)) and zero (\(z_i\)) and the gain (\(k\)) of the IIR filter. The constants were re-substituted in the transfer function which was subsequently inverted. The research findings have showed that the implementation of the proposed scheme could address the attenuation phenomenon, compensate the system bandwidth and enhance the time response of the output signal.
2. Conventional Design

Figure 1 illustrates a typical digital signal processing (DSP) system in which the input filter is of low pass filter whose function is to prevent the high frequency signal input entering the system, and the output filter is also of low pass filter that transforms the staircase signal from the D/A converter into a continuous signal. In theory, the cut-off frequency \( \omega_c \) of the output filter is \( f_s/2 \) (i.e. the Nyquist frequency) \([1-6]\). The DSP bandwidth is thus governed by the cut-off frequency of the analog output filter, e.g. for a digital signal processing system operating at the 12.8kHz sampling frequency \( f_s \), the cut-off frequency of the analog output filter is half of \( f_s \) or 6.4kHz. Figure 2 depicts the bandwidth of a conventional DSP system which suffers from the attenuation phenomenon which is more pronounced approaching the Nyquist frequency.

![Fig. 2 Bandwidth of conventional digital signal processing system](image)

In \([8]\), the design of an IIR digital filter in the DSP system using the approximation of derivative method was proposed, whose cut-off frequency is equal to that of the analog output filter. The IIR transfer function was inverted and then converted into a difference equation for bandwidth compensation. Despite the bandwidth enhancement, the technique suffers from the discrepancy in the magnitude responses of the IIR digital filter and the analog output filter, a phenomenon attributable to the fact that the development of the IIR digital filter was predicated on the approximation of derivative technique. For instance, Fig. 3 compares the magnitude responses of the first-order analog output filter and the IIR filter at the 10, 12 and 14 kHz sampling frequencies, given that the cut-off frequency is 2kHz.

![Fig. 3 Magnitude responses of the first-order IIR digital filter based on the approximation of derivative technique at Fs of 10, 12 and 14kHz vis-à-vis that of the output filter](image)

Interestingly, at the 14kHz sampling frequency, the magnitude response of the IIR filter increasingly resembles that of the output filter, indicating that the higher sampling frequency, the more resembling the IIR filter magnitude response to the analog output filter’s. Notwithstanding, considerable computing resources are required for the high sampling frequency.

3. Proposed Digital Signal Processing System

The bandwidth compensation performance in a digital signal processing system is positively correlated to the degree of resemblance between the magnitude responses of the IIR filter and the analog output filter. In fact, the magnitude responses belonging to the IIR and analog output filters differ due to the warping frequency or the nonlinear relationship between the analog and digital frequencies. The warping frequency is attributable to the utilization of the approximation of derivative technique or the bilinear transform technique \([1,2]\) and contributes to the discrepancy in the magnitude responses of the IIR and analog output filters. To address the discrepancy issue and thus enhance the compensation performance, it is necessary to keep the warping frequency incidence to a minimum.

The proposed scheme aims to achieve the magnitude response gains of the IIR filter at the 0, \( \omega_c \) and \( \pi \) frequencies that are identical to those of the analog output filter. The respective magnitude response gains of both filters are thus 1, \( 1/\sqrt{2} \) and \( c_1 \) \([9,10]\), as shown in Fig. 4 and Table 1. The sameness renders the relationship between the analog and digital frequencies linear. The parameter \( c_1 \) (i.e. the magnitude response gain at the Nyquist frequency) and the transfer function of the IIR filter using the pole-zero placement technique are respectively expressed in Eqs. (1) and (2).

![Fig. 4](image)

| Frequency (Hz) | Magnitude (dB) |
|---------------|---------------|
| 0             | 1             |
| \( \omega_c \) | \( 1/\sqrt{2} \) |
| \( \pi \)     | \( c_1 \)     |

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Fig. 4 Magnitude response of first order analog output filter

Table 1 Gain of IIR compensation filter at frequencies 0, \( \omega_c \) and \( \pi \)

| Frequency \((\omega)\) | Gain |
|-----------------------|------|
| 0                     | 1    |
| \( \omega_c \)        | \( \frac{1}{\sqrt{2}} \) |
| \( \pi \)             | \( c_1 \) |

\[
c_1 = \frac{1}{\sqrt{1 + \frac{f_n^2}{f_c^2}}} \quad (1)
\]

where \( c_1 \) is the magnitude response gain at the Nyquist frequency, \( f_n \) is the Nyquist frequency and \( f_c \) is the cut off frequency.

\[
H(z) = k \left( \frac{z - z_1}{z - p_1} \right) \quad (2)
\]

Substitute \( z \) in Eq. (2) with \( z = e^{j\omega} \) to derive Eq. (3).

\[
H(\omega) = k \left( \frac{e^{j\omega} - z_1}{e^{j\omega} - p_1} \right) \quad (3)
\]

At \( \omega = 0 \) where the magnitude response gain is 1, substitute the values in Eq. (3) to obtain Eq. (4) which is then rearranged into Eq. (5).

\[
H(0) = k \left( \frac{e^{j0} - z_1}{e^{j0} - p_1} \right)
\]
\[
1 = k \left( \frac{e^{j0} - z_1}{e^{j0} - p_1} \right)
\]
\[
1 - p_1 = k \left( 1 - z_1 \right) \quad (5)
\]

At \( \omega = \pi \) where the magnitude response gain is \( c_1 \), substitute the values in Eq. (3) to obtain Eq. (6) which is then rearranged into Eq. (7). Meanwhile, at \( \omega = \omega_c \) where the magnitude response gain is \( 1/\sqrt{2} \), substitute the values in Eq. (3) to obtain Eq. (8) which is then rearranged into Eq. (9) and then Eq. (10) until arriving at Eq. (11).

The parameters \( k, p_1 \) and \( z_1 \) are concurrently solved using Eqs. (5), (7) and (11). Since these equations are nonlinear, the Newton-Raphson method \([11,12]\) is thus used and expressed in Eq. (12).

\[
1 = k \left( \frac{e^{j\pi} - z_1}{e^{j\pi} - p_1} \right) \quad (6)
\]

\[
c_1 = k \left( \frac{e^{j\omega} - z_1}{e^{j\omega} - p_1} \right) \quad (7)
\]

\[
H(\omega_c) = k \left( \frac{e^{j\omega_c} - z_1}{e^{j\omega_c} - p_1} \right) \quad (8)
\]

\[
\frac{1}{\sqrt{2}} = k \left( \frac{e^{j\pi} - z_1}{e^{j\pi} - p_1} \right) \quad (9)
\]

Define \( e^{j\omega} = \cos(\omega) + j\sin(\omega) \)

\[
\frac{1}{\sqrt{2}} = k \left( \frac{\cos(\omega) + j\sin(\omega) - z_1}{\cos(\omega) + j\sin(\omega) - p_1} \right) \quad (10)
\]
\[
x_{n+1} = x_n - J(k_n,p_{in},z_{in})^{-1} f(k_n,p_{in},z_{in}) \tag{12}
\]

where
\[
x_n = \begin{bmatrix} k_n \\ p_{in} \\ z_{in} \end{bmatrix}, f(k_n,p_{in},z_{in}) = \begin{bmatrix} f_1(k_n,p_{in},z_{in}) \\ f_2(k_n,p_{in},z_{in}) \\ f_3(k_n,p_{in},z_{in}) \end{bmatrix}, J(k_n,p_{in},z_{in}) = \begin{bmatrix} \frac{\partial f_1}{\partial k} & \frac{\partial f_1}{\partial p_{in}} & \frac{\partial f_1}{\partial z_{in}} \\ \frac{\partial f_2}{\partial k} & \frac{\partial f_2}{\partial p_{in}} & \frac{\partial f_2}{\partial z_{in}} \\ \frac{\partial f_3}{\partial k} & \frac{\partial f_3}{\partial p_{in}} & \frac{\partial f_3}{\partial z_{in}} \end{bmatrix}
\]

\[
f_1 = k_n(1-z_{in}) - (1-p_{in}), f_2 = k_n(1+z_{in}) - c_1(1+p_{in})
\]

and
\[
f_3 = k_n^2 \left( \frac{1-2p_{in}\cos(\omega_c) + p_{in}^2 + (\cos(\omega_c) - p_{in})(p_{in} - z_{in})^2 + (z_{in}\sin(\omega_c) - p_{in}\sin(\omega_c))^2}{1-2p_{in}\cos(\omega_c) + p_{in}^2} \right)
\]

In Eq. (12), given that the iteration number \(n\) is 1, the vector \(x_i\) is then the initial value within the -1 to 1 interval inside the unit circle. The iteration is repeated until the convergence of \(k\), \(p_1\) and \(z_1\) are achieved and subsequently substituted in Eq. (2) to derive the transfer function of the IIR filter. To arrive at the bandwidth compensatory scheme, the transfer function is inverted and then converted into the difference equation, as respectively presented by Eqs. (13) and (14). Figure 5 illustrates the digital signal processing system integrated with the proposed bandwidth compensatory scheme.

\[
H_i(z) = \frac{1}{k} \left( \frac{z-p_1}{z-z_1} \right) \tag{13}
\]

\[
y(n) = \frac{1}{k} x(n) - \frac{1}{k} p_1 x(n-1) + z_1 y(n-1) \tag{14}
\]

where \(x(n)\) and \(y(n)\) are respectively the input and output of the compensated DSP system.

As an example, given the cutoff \((\omega_c)\) and sampling frequencies of 2 kHz and 14 kHz, respectively, the Nyquist frequency \((\pi)\) is 7 kHz and \(c_1\) and \(\omega_c\), as respectively calculated by Eqs. (1) and (15), are 0.2747211 and 0.8975979. Substitute \(c_1\) and \(\omega_c\) into Eqs. (5), (7) and (11) for \(k = 0.4976400\), \(p_1 = 0.3852877\) and \(z_1 = 0.2352548\) which are subsequently substituted in Eq. (2) for the IIR filter transfer function (Eq. (16)). Figure 6 compares the magnitude responses of the analog output filter and the IIR filter integrated with the proposed compensatory scheme.

\[
\omega_c = \frac{2\pi f_s}{f_s} \tag{15}
\]

where \(f_s\) is the sampling frequency.

\[
H(z) = 0.4976400 \left( \frac{z+0.2352548}{z-0.3852877} \right)
\]

\[
H(z) = 0.4976400 \left( \frac{1+0.2352548z^{-1}}{1-0.3852877z^{-1}} \right) \tag{16}
\]

Fig. 5 Digital signal processing system integrated with the proposed bandwidth compensatory scheme.
4. Simulation and Experimental Results

To verify the effectiveness with regard to the bandwidth compensation of the proposed scheme, simulations were carried out using MATLAB. In the simulation, the cut-off frequencies were either 1.5kHz or 2 kHz, while the sampling frequencies were 10 kHz, 12 kHz and 14 kHz. Table 2 tabulates the simulated $k$, $z_i$, $p_i$ and the IIR filter transfer function $s$ under investigation. In Figs. 7(a)-(f), comparisons are made between the simulated magnitude response of the proposed scheme, that of the analog output filter and that based on the approximation of derivative for the two cut-off and three sampling frequencies under study. Table 3 compares the simulated magnitude response gains of the analog output filter and IIR filter at the 0, $\omega_c$ and $\pi$ frequencies.

To further validate its bandwidth compensatory capability, the proposed scheme was experimented using a TMS320C31 DSP starter kit [13-16] connected to the MAX-547 12-bit external D/A interface board which is connected to two first order analog output filters. The A/D converter is of 14-bit resolution (TLC32040) and the digital signal processor is of TMS320C31 operating at the 21.853-kHz sampling frequency. In addition, the Tektronix TDS-3014 digital oscilloscope and the Tektronix AFG 320 arbitrary signal generator were utilized in the experiments. Due to the idiosyncratic characteristics of the TLC32040 A/D converter, the sampling frequency is not a whole number. Moreover, in the experiments, the maximum capacity of the A/D converter (i.e. 21.853 kHz) was deliberately used as the sampling frequency.

In Fig. 8, the input signal ($x(t)$), the output signals of D/A MAX547 in the absence of the analog output filter, in the presence of the analog output filter, and in the presence of the analog output filter with the bandwidth compensatory scheme, using the oscilloscope, are respectively represented by channels (CH) 1-4. The unity gain amplifier software program was utilized in the experiments. In determination of the magnitude response, two cut-off frequencies of 1.5kHz and 2kHz were used and the input signal frequency was varied from 50Hz, 500Hz, 1kHz, 1.5kHz, 3kHz to 5kHz. Figures 9(a)-(f) illustrate the input signal ($x(t)$), the output signals in the absence and presence of the analog output filter, and in the presence of the analog output filter with the bandwidth compensatory scheme for the 1.5kHz cut-off frequency for the various input signal frequencies. Meanwhile, Fig. 11 compares the magnitude responses of the non-compensated and compensated output filters at the 1.5kHz and 2kHz cut-off frequencies.

![Fig. 8 Comparison of the input signal (CH1), the output signals in the absence (CH2) and presence (CH3) of the analog output filter and in the presence of the analog output filter with the bandwidth compensatory scheme (CH4)](image)
### Table 2 Simulated $k$, $p_i$, $z_i$ and the IIR filter transfer functions, $H(z)$, at the cut-off and sampling frequencies under investigation

| Cut-off frequency | Sampling frequency | $k$         | $p_i$       | $z_i$       | $H(z)$             |
|------------------|-------------------|------------|------------|------------|-------------------|
| 1.5kHz           | 10kHz             | 0.51373309| 0.36466839| -0.23669589| $H(z) = \frac{1+0.23669589z^{-1}}{1-0.36466839z^{-1}}$ |
|                  | 12kHz             | 0.45470299| 0.43979577| -0.23202228| $H(z) = \frac{1+0.23202228z^{-1}}{1-0.43979577z^{-1}}$ |
|                  | 14kHz             | 0.40753197| 0.49902550| -0.22928882| $H(z) = \frac{1+0.22928882z^{-1}}{1-0.49902550z^{-1}}$ |
| 2kHz             | 10kHz             | 0.61155251| 0.23589475| -0.24945155| $H(z) = \frac{1+0.24945155z^{-1}}{1-0.23589475z^{-1}}$ |
|                  | 12kHz             | 0.54908122| 0.31891514| -0.2404082 | $H(z) = \frac{1+0.2404082z^{-1}}{1-0.31891514z^{-1}}$ |
|                  | 14kHz             | 0.49764004| 0.38528771| -0.23525486| $H(z) = \frac{1+0.23525486z^{-1}}{1-0.38528771z^{-1}}$ |

### Table 3 Simulated magnitude response gains of the analog and IIR filters at the $0$, $\omega_c$ and $\pi$ frequencies

| Cut-off frequency | Analog filter | IIR Digital filter with proposed design |
|------------------|--------------|----------------------------------------|
|                  | Gain at frequencies | Gain at frequencies | Gain at frequencies |
|                  | $f_c$ / 2 | $\omega_c$ | $\pi$ | $f_c$ / 2 | $\omega_c$ | $\pi$ |
| 1.5kHz           | 10kHz       | 0.70710678118655 | 0.28734788556635 | 0.707181730704827 | 0.287347919701994 |
|                  | 12kHz       | 0.70710678118655 | 0.24253562503633 | 0.707168132255925 | 0.242535645161976 |
|                  | 14kHz       | 0.70710678118655 | 0.20952908873087 | 0.707158805634759 | 0.209529101558967 |
| 2kHz             | 10kHz       | 0.70710678118655 | 0.37139067635410 | 0.707185249255396 | 0.371390720121469 |
|                  | 12kHz       | 0.70710678118655 | 0.31622776061684 | 0.707170108198330 | 0.316227792016905 |
|                  | 14kHz       | 0.70710678118655 | 0.27472112789738 | 0.707160027764203 | 0.274721144576873 |

### Table 4 Rise time and fall time at the 1.5kHz and 2kHz cut-off frequencies for various pulse widths

| Pulse width | Rise time (µs) | Fall time (µs) | Rise time (µs) | Fall time (µs) | Rise time (µs) | Fall time (µs) | Rise time (µs) | Fall time (µs) | Rise time (µs) | Fall time (µs) |
|-------------|----------------|----------------|----------------|----------------|----------------|----------------|----------------|----------------|----------------|----------------|
| Without the analog output filter | With the analog output filter | Propose design | Without the analog output filter | With the analog output filter | Propose design |
| 10 ms       | 32.70          | 39.25          | 248.1          | 248.8          | 42.60          | 43.43          | 203.4          | 211.9          | 39.40          | 41.92          |
| 5 ms        | 37.23          | 38.21          | 249.0          | 246.9          | 41.35          | 41.54          | 215.7          | 199.8          | 40.20          | 41.16          |
| 1 ms        | 37.55          | 37.73          | 241.8          | 245.8          | 39.93          | 40.68          | 204.6          | 200.1          | 39.70          | 39.73          |
| 500 µs      | 33.67          | 38.37          | 217.6          | 218.6          | 39.50          | 39.50          | 189.7          | 177.3          | 38.98          | 38.51          |
Fig. 7  Simulated magnitude responses of the approximate of derivative-based scheme, the proposed schemes and the analog output filter
Fig. 9 Input and output signals at the 1.5 kHz cut-off frequency for the various input signal frequencies of 50Hz, 500Hz, 1kHz, 1.5kHz, 3kHz and 5kHz, where CH1, CH2, CH3 and CH4 respectively represent the input signal, the output signals in the absence and presence the analog output filter and in the presence of the analog output filter with the compensatory scheme.
Fig. 10 Input and output signals at the 2 kHz cut-off frequency for the various input signal frequencies of 50Hz, 500Hz, 1kHz, 1.5kHz, 3kHz and 5kHz, where CH1, CH2, CH3 and CH4 respectively represent the input signal, the output signals in the absence and presence the analog output filter and in the presence of the analog output filter with the compensatory scheme.
Figures 12(a)-(d) respectively illustrate the time responses at the 1.5kHz cut-off frequency for the square wave input signal with the pulse widths of 10ms, 5ms, 1ms and 100μs, while Figs. 13(a)-(d) depict those at the 2kHz cut-off frequency under the same conditions. Table 4 compares the rise time and fall time at both cut-off frequencies for the various pulse widths under study.

In this research, the measurements of total harmonic distortion (THD) [17] were taken using a dynamic signal analyzer Agilent 35670A at the frequency range of 50Hz-5kHz. Table 5 tabulates the THD (%) from the fundamental frequency through to 10th harmonic frequencies for the frequency range under study.
Fig. 13  Time response at the 2kHz cut-off frequency for the various pulse widths of 10ms, 5ms, 1ms and 500µs, where CH1, CH2, CH3 and CH4 respectively represent the input signal, the output signals in the absence and presence the analog output filter and in the presence of the analog output filter with the compensatory scheme.

Table 5  Total harmonic distortion (THD) of the output signal

| Frequency (Hz) | Signal generator (THD) | D/A output without the analog output filter | The 1.5kHz cut-off frequency of analog output filter | The 2kHz cut-off frequency of analog output filter |
|---------------|------------------------|--------------------------------------------|--------------------------------------------------|--------------------------------------------------|
|               |                        |                                            | Without compensation | With compensation | Without compensation | With compensation |
| 50            | 0.0312                 | 0.0731                                     | 0.0730               | 0.0680             | 0.0749               | 0.0646             |
| 100           | 0.0304                 | 0.0751                                     | 0.0760               | 0.0710             | 0.0744               | 0.0645             |
| 200           | 0.0305                 | 0.0758                                     | 0.0630               | 0.0630             | 0.0682               | 0.0595             |
| 500           | 0.0304                 | 0.0784                                     | 0.0460               | 0.0490             | 0.0547               | 0.0480             |
| 1000          | 0.0304                 | 0.0805                                     | 0.0350               | 0.0350             | 0.0375               | 0.0463             |
| 1500          | 0.0305                 | 0.0864                                     | 0.0330               | 0.0350             | 0.0355               | 0.0459             |
| 2000          | 0.0305                 | 0.0891                                     | 0.0360               | 0.0320             | 0.0325               | 0.0423             |
| 2500          | 0.0305                 | 0.0990                                     | 0.0430               | 0.0350             | 0.0313               | 0.0383             |
| 3000          | 0.0308                 | 0.1290                                     | 0.0490               | 0.0400             | 0.0335               | 0.0396             |
| 3500          | 0.0308                 | 0.1058                                     | 0.0500               | 0.0420             | 0.0458               | 0.0409             |
| 4000          | 0.0309                 | 0.1089                                     | 0.0600               | 0.0550             | 0.0463               | 0.0471             |
| 4500          | 0.0312                 | 0.1142                                     | 0.0670               | 0.0750             | 0.0501               | 0.0671             |
| 5000          | 0.0316                 | 0.2282                                     | 0.0770               | 0.0950             | 0.0655               | 0.0873             |
| **Average**   | 0.0307                 | 0.1033                                     | 0.0544               | 0.0534             | 0.0500               | 0.0531             |
5. Conclusion

This research has proposed the novel algorithmic scheme to tackle the bandwidth attenuation problem commonly found in the conventional digital signal processing (DSP) system. The scheme development started with the determination of the analog output filter gains at the zero, cut-off \( (\omega_c) \) and Nyquist \( (f_s/2) \) frequencies and then their respective nonlinear system equations. The Newton-Raphson method was subsequently applied for the positions of pole \( (p_1) \) and zero \( (z_1) \) and the gain \( (k) \) of the IIR filter. The constants were re-substituted in the initial transfer function and an inversion carried out. The aim of the scheme implementation is to achieve the identical magnitude responses of the target (compensation) filter and the conventional analog output filter. To verify the proposed scheme, the simulations were carried out at the two cut-off frequencies (i.e. 1.5 and 2kHz) for the three sampling frequencies of 10, 12 and 14kHz. The simulation results demonstrated that the magnitude responses of the compensation (IIR) and conventional analog output filters are more in sync than are those of the approximation of derivative-based and conventional filters. To further validate the effectiveness of the scheme, the experiments were undertaken and the findings showed that the initially staircase output signal becomes flat (i.e. continuous) and non-attenuating and the bandwidth compensated. Moreover, the time response of the output signal under the proposed scheme is considerably shorter relative to that of the non-compensated output filter. It can thus be concluded that the proposed algorithmic scheme is both highly operationally efficient and cost-effective in tackling the attenuation problem through the bandwidth compensation.

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Rodjarin Boontawan received her B.Sc. degree in physics of science and her M.Sc. degree in medical instrumentation of industrial physics and medical instrumentation of applied science from the King Mongkut’s University of Technology North Bangkok, Bangkok, Thailand, in 2003 and 2006, respectively. She has studied in the doctoral degree program in electrical engineering at King Mongkut’s Institute of Technology Ladkrabang, Bangkok, Thailand. Her research interests are digital filter and application.

Siraphop Tooprakai received his B.S. in materials science from Chiang Mai University, Chiang Mai, Thailand, in 1995 and his M. Eng. and D. Eng. degrees in electrical engineering from the King Mongkut’s Institute of Technology Ladkrabang (KMITL), Bangkok, Thailand, in 1998 and 2008, respectively. Since 2007, he has been a member of the Department of Telecommunication Engineering at Faculty of Engineering, KMITL, where he is currently an Assistant Professor of telecommunication. His research interests are in the digital circuit design, communication network and communication circuit design.

Kobchai Dejhan received his B. Eng. and M. Eng. degrees in electrical engineering from the King Mongkut’s Institute of Technology Ladkrabang (KMITL), Bangkok, Thailand in 1978 and 1980, respectively, and his Doctoral degree in telecommunications from the Ecole Nationale Superieure des Telecommunications (ENST) Paris, France (Telecom Paris) in 1989. Since 1980, he has been a member of the Department of Telecommunication at Faculty of Engineering, KMITL. He is an Associate Professor. He also was with the Research Center for Communication and Information Technology (ReCCIT) at KMITL as a Chief of Communications Circuit Designs Laboratory. His research interests are in the area of communication circuit designs, signal processing, VLSI and CMOS integrated circuit design. He is a member of IEICE, Japan, and a senior member of IEEE, USA.

Surapun Yimman received his M. Eng. degree from the King Mongkut’s Institute of Technology North Bangkok, Bangkok, and D. Eng. degree from the King Mongkut’s Institute of Technology Ladkrabang, Bangkok, all in electrical engineering in 1999 and 2007, respectively. He is an Associate Professor of Department of Industrial Physics and Medical Instrument at Faculty of Applied Science at the King Mongkut’s University of Technology North Bangkok, Bangkok. His research interests are in the digital signal processing and applications.

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