Results and discussion: System performance is demonstrated with examples of recursive and non-recursive circuits of varying complexity. The circuits are an eighth order Avenhaus filter [5], a second order IIR lattice filter, a three tap FIR filter and an eight tap FIR filter. Fig. 3 illustrates the DFG obtained for the three tap FIR filter before the application of loop unfolding. The shortest critical path has been achieved through pipelining, as in this example a multiplication has an execution time twice that of an addition.

The results obtained for each of the example circuits are presented in Table 1. The comparative power is an estimation of the power consumption of the new circuit expressed as a percentage of the original power consumption. In all cases power was reduced.

Conclusions: The results demonstrate that the GA is capable of optimising circuits for power at a high level using a number of transformations. The variable application rates produced designs with good area/power tradeoff. The concurrent tracking of power, area and speed shows the effectiveness of the GA method.

References

Fig. 3 Three tap FIR filter

Fig. 4 GA performance with eighth order Avenhaus filter

Table 1: Results for the example circuits

<table>
<thead>
<tr>
<th>Circuit</th>
<th>Percentage power</th>
</tr>
</thead>
<tbody>
<tr>
<td>Eighth order Avenhaus</td>
<td>Before unfold: 28</td>
</tr>
<tr>
<td>Second order lattice</td>
<td>65</td>
</tr>
<tr>
<td>Three tap FIR</td>
<td>44.9</td>
</tr>
<tr>
<td>Eight tap FIR</td>
<td>20.54</td>
</tr>
</tbody>
</table>

Synchronous dimming control for a cold-cathode fluorescent lamp driver

M.-S. Lin, M.-C. Lee, D.Y. Chen and W.-S. Feng

Indexing terms: Power electronics, Fluorescent lamps

A novel synchronous dimming control circuit is proposed and tested for a cold-cathode fluorescent lamp (CCFL) driver circuit. The scheme is simple and efficient, and it alleviates problems associated with lamp flickering and EMI control.

Introduction: Cold-cathode fluorescent lamps (CCFLs) are widely used to illuminate the liquid crystal displays (LCDs) used in many battery powered instruments. In such an application, the driver circuits to CCFLs must be able to provide stable power with efficiency. In addition, a dimming control capability is a very desirable feature. Fig. 1 shows a typical driver circuit which uses a parallel-resonant inverter circuit regulated by a Buck converter. The dimming control is accomplished by using the pulse width modulated (PWM) controller to regulate the power to the lamps.
The lamp illumination is roughly proportional to the lamp current \([4, 5]\) so the lamp current is used as a feedback signal in this circuit. There are, however, two drawbacks of this circuit. The power circuit is the same as the conventional circuit shown in Fig. 1, but the dimming control is accomplished with a comparator circuit and an additional feedback signal derived from an inverter circuit at point \(A\).

**Power circuit operation:** Referring to Fig. 3, the circuit starts to operate as the current flows through \(R_1\) and \(Q_1\), and \(Q_2\) is turned on. Then the capacitor \(C\) and the primary magnetising inductance of the transformer \(T_1\) begin to resonate with frequency \(1/(2\pi\sqrt{L_C}C)\), where \(L_C\) is the total primary magnetising inductance of transformer \(T_1\). The polarity of the auxiliary winding \(N_A\) keeps the switch \(Q_1\) ON and the switch \(Q_2\) OFF. The voltage at point \(A\) is equal to \(-v_C/2\) and the voltage of the capacitor \(C\) is negative during this half-cycle. When the polarity of the capacitor voltage is reversed owing to resonance, \(Q_1\) turns on and \(Q_2\) turns off. The voltage at point \(A\) is equal to \(v_C/2\) and the voltage of the capacitor \(C\) is positive during this half-cycle. As the resonance continues, \(Q_1\) and \(Q_2\) switch alternatively. It can be seen that the voltage at point \(A\) is equal to \(v_C/2\) and has a frequency that is double the resonant frequency of inverter circuit. This signal will be used for dimming control, to be described in the next section.

**Dimming control circuit:** Referring again to Fig. 3, the voltage of the non-inverting terminal of the comparator consists of two components. One component is a DC signal derived from the DC voltage across \(C_1\) and the other component is an AC signal derived from the voltage at point \(A\). Fig. 4 shows the sketched waveforms of the non-inverting terminal \(V_+\) and the output voltage \(V_o\) of the comparator. Note that the switching frequency of \(V_o\) (which is used to drive the Buck converter) is determined by the AC component of \(V_+\), which is twice the resonant frequency of inverter circuit. In this way, the Buck converter switching frequency is always synchronous with the inverter resonant frequency, thereby avoiding the beat frequency. Dimming control is accomplished by simply varying \(V_o\). Referring to Fig. 4, as \(V_o\) is varied the duty cycle of Buck converter switching is varied and the lamp current is varied. If the DC component of \(V_o\) is made much larger than the AC component, then the lamp current is controlled by \(V_o\). Therefore the lamp illumination can be controlled by varying \(V_o\).

### Determination of the \(R_2\), \(R_3\), and \(R_4\) values:
The DC component of the non-inverting terminal of the comparator (in Fig. 3) is
\[
V_{+_{DC}} \approx V_{ref} = \frac{R_2}{R_3 + R_4} \tag{1}
\]
and the AC voltage component is
\[
V_{+_{AC}} = \frac{R_A}{R_3 + R_4} \tag{2}
\]
From eqns. 1 and 2 and the fact that the DC component is designed to be much larger than the AC component (so that lamp illumination is controlled by \(V_{ref}\), the \(R_2\), \(R_3\), \(R_4\) values should then satisfy the following inequalities:
\[
R_2/R_3 \gg R_4 \tag{3}
\]
\[
R_3/R_4 \ll R_2 \tag{4}
\]
Table 1: Part list

<table>
<thead>
<tr>
<th>L</th>
<th>C</th>
<th>N_c</th>
<th>N_f</th>
<th>R_c</th>
<th>R_f</th>
<th>C_c</th>
<th>C_f</th>
<th>R_c</th>
<th>R_f</th>
</tr>
</thead>
<tbody>
<tr>
<td>400μH</td>
<td>0.2μF</td>
<td>16</td>
<td>1300</td>
<td>1kΩ</td>
<td>1kΩ</td>
<td>4.7μF</td>
<td>0.1μF</td>
<td>20μΩ</td>
<td>10μΩ</td>
</tr>
</tbody>
</table>

Experimental verification: A 2.5W driver circuit has been built using the parts listed in Table 1. Fig. 3 shows the lamp voltage waveform. Note that there is no beat frequency, which may eliminate one reason for the flickering lamp problem. In addition, it is good for EMI control.

Fig. 5 Lamp voltage waveform measured from the circuit of Fig. 3. Note that there is no beat frequency.

The power consumption of the comparator is 12mW (as compared with 150mW for a PWM control IC). This improves the efficiency by about 5% for a 2.5W CCFL driver circuit. In addition, a PWM controller costs about US$0.46 and a comparator costs about $0.15. For a 2.5W CCFL driver this proposed scheme reduces the part cost by 9%.

Conclusions: A novel synchronous dimming control driver circuit has been proposed. It features high efficiency, low cost, simplicity and ease of EMI control. It also makes the lamp drive a single frequency signal, thus eliminating a cause of the lamp flickering encountered in a conventional circuit. The circuit has been experimentally verified in a 2.5W driver circuit.

© IEE 1996 15 March 1996
Electronics Letters Online No: 19960779
M.-S. Lin, M-C. Lee and W.-S. Feng (Department of Electrical Engineering, National Taiwan University, Taipei, Taiwan 106, Republic of China)
D.Y. Chen (Department of Electrical Engineering, Virginia Polytechnic Institute and State University, Blacksburg, VA 24061, USA)

References
3 Williams, J.: ‘Fluorescent lamp power supply and control unit’, US patent, P/N 5,408,162, Apr. 18, 1995
5 Williams, J.: ‘Techniques illuminate backlit LCDs with high efficiency’, EDN Mag., 1994, pp. 89–90

Adaptive predistortion with reduced feedback complexity

A.R. Mansell and A. Bateman

Indexing terms: Radiofrequency amplifiers, Power amplifiers

A technique for the adaption of a complex-gain baseband predistorter is proposed, using only a single coherent downconversion in the feedback path. Despite this reduction in complexity when compared to previous implementations, the adaption process can still correct for both AM/AM and AM/PM nonlinearities in the RF power amplifier. Practical results from a 220MHz transmitter are given.

Introduction: The use of digital modulation formats that use variations in both carrier amplitude and carrier phase has led to the development of linear wireless transceiver architectures. Adaptive baseband predistortion [1, 2] has been proposed as a suitable RF power amplifier linearisation technique for such transceivers. Its use of discontinuous modulation feedback to track the nonlinear power amplifier characteristics removes the stability problems experienced with continuous modulation feedback techniques such as the Cartesian loop [3]. It is, however, more susceptible to non-ideal component behaviour in the transmitter RF hardware [4]. A technique is proposed which does not require a quadrature demodulator in the feedback path; this leads to a reduction in complexity and removes one possible source of performance degradation.

Fig. 1 Complex-gain predistortion transmitter

Theory: Fig. 1 shows the basic architecture of an adaptive complex-gain predistorter. The look-up table (LUT) is indexed by some function of the input signal power, and it stores a piecewise approximation to the required gain predistortion characteristic. Traditionally, the adaption algorithm has used both the input modulating signal (x_i, x_q) and the feedback modulation (i_f, q_f) derived by a quadrature demodulator, to estimate the error in the LUT entries.

Assuming that, at some input power level P, the nonlinear power amplifier has a complex-envelope gain of (a+jb) then it is trivial to show that at this point

\[ i_f = a_i - b q_i \]  

If we now take two data sets, \((i_x, q_x, i'_x, q'_x)\) and \((i_y, q_y, i'_y, q'_y)\), both of which represent vectors that fall within some small power range (P±5dB) then eqn. 1 can be used to show

\[ \lim_{\beta \to a} = \frac{i'_f q_y - i'_x q_y}{i'_x q_y - i'_x q_y} \]  

\[ \lim_{\beta \to b} = \frac{i'_x q_y - i'_x q_y}{i'_x q_y - i'_x q_y} \]  

Hence it is possible to estimate the complex-envelope gain of the power amplifier chain, at a particular operating point, without knowledge of \( q_f \); consequently, a single mixer and ADC may be used in the feedback path instead of the full quadrature demodulation (Fig. 2).

Practical deployment of this technique requires that the variation in power amplifier gain across any band of power of width