A COMPARISON OF CAP/QAM ARCHITECTURES

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ABSTRACT

Recently CAP modulation (Carrierless Amplitude and Phase modulation) has been of wide interest to industry because of its simplicity, bandwidth efficiency and zero dc component [2-5]. This paper compares CAP versus QAM modulation and highlights some practical implementation issues. In addition, three different CAP-like architectures are compared according to A/D sampling-rates, A/D resolutions and jitter requirements.

1. INTRODUCTION

CAP is a modulation scheme in digital communication systems and can be considered as a bandpass PAM (Pulse Amplitude Modulation) [1], in which the carrier frequency is near baseband. On the other hand it can also be viewed as a variation of QAM (Quadrature Amplitude Modulation) without explicit modulation/demodulation blocks. In contrast to uncoded PAM, CAP has zero dc power. This spectral characteristic makes it well suited to ac coupled channels. It also allows the possibility of leaving a frequency band, near dc, which can be used for other signals (such as standard telephone signals).

Most CAP systems are designed and implemented in the digital domain by applying a relatively high-rate and high-resolution A/D thereby taking advantage of digital signal processing. However, as we shall see, using an analog CAP demodulator allows the use of less-costly A/D converters. Also, we shall see that CAP systems suffer from higher sampling jitter sensitivity.

This paper compares CAP and QAM in detail as well as presenting various architectures which can be considered for different applications.

2. CAP VERSUS QAM

2.1. QAM

Fig. 1 shows a simple QAM system in which $g(t)$ and $f(t)$ are lowpass shaping filters. Since we are going to compare QAM and CAP systems, $\omega_c$ represents a carrier frequency near baseband which is equal or greater than the bandwidth of $g(t)$ (i.e. $F_c \geq F_g/2$). In addition, the carrier signals in both the transmitter and receiver must be in phase.

In Fig. 1, $A_k$ and $B_k$ are the input symbols and $y_A(t)$, $y_B(t)$ are the output signals of demodulators in the receiver which contain the transmitted symbol information. Although the inphase and quadrature signals in the transmitter are summed together, we wish the least interference between the two outputs of the demodulator in the receiver.

The output of receiver filters is divided into two components of inphase or quadrature ($y_I(t)$ or $y_Q(t)$) and the interference signal ($y_{IQ}(t)$ or $y_{QI}(t)$):

$$y_{A}(t) = y_{I}(t) + y_{IQ}(t)$$

(1)

Regarding Fig. 1, these components can be written as:

$$y_{I}(t) = \left( \left[ \sum_{k} A_k g(t-kT) \right] \cos(\omega_c t)^2 \right) \otimes f(t)$$

(2)

or:

$$y_{I}(t) = \sum_{k} A_k h(t-kT) +$$

$$0.5 \left[ \sum_{k} A_k g(t-kT) \cos 2(\omega_c t) \right] \otimes f(t)$$

(3)

in which:

$$h(t) = 0.5(g(t) \otimes f(t))$$

(4)

and:

$$y_{IQ}(t) = \left( 0.5 \sum_{k} B_k g(t-kT) \sin 2(\omega_c t) \right) \otimes f(t)$$

(5)

Since $f_c$ is greater than the bandwidth of the lowpass shaping filters $g(t)$ and $f(t)$, the second term in (3) and

![Figure 1: QAM Modulation Scheme: Transmitter and Receiver](image)

1. "\otimes" is the convolution sign and $T$ is the symbol period
also $y_{IQ}(t)$ is negligible for all $t$ (not only on sampling points). So we can write:

$$y_A(t) = \sum_k A_k h(t - kT)$$

and in a similar way it can be seen that:

$$y_B(t) = \sum_k B_k h(t - kT)$$

Fig. 2 shows the demodulator’s output; $y_A(t)$, and its components $y_A(t)$ and $y_{IQ}(t)$ for a single complex symbol. It should be noted that in this paper, all examples make use of root-raised cosine pulses with 20% excess bandwidth for $g(t)$ and $f(t)$ (so $h(t)$ is nyquist) and whenever necessary a 50MHz baud-rate has been used.

As we expected the interference between I and Q signals is near zero and in this simple example, ignoring the channel effects, just by sampling the output of demodulator, the transmitted symbols can be recovered.

### 2.2. CAP

Considering Fig. 1, one can imagine a system in which instead of a lowpass filter followed by a multiplier, a bandpass filter with relevant centre frequency and bandwidth be used.

Fig. 3 shows this idea in which I and Q filters have 90 degrees difference in phase. Such a system is called CAP or Carrierless AM/PM. It should be noted that this idea is practical whenever the carrier frequency is not very high, so we can have reasonable filters. For example in the digital domain we can approximate those filters by reasonable number of taps for low $f_c$, whereas it would not be feasible for high values of $f_c$.

At first glance, the systems in Fig. 1 and Fig. 3 may seem the same, but that is not the case. The Modulation-De-modulation portions in Fig. 1 are linear and time variant systems but in Fig 3 they are both linear and time invariant filters. This observation has important consequences for the signal between sampling instances. More precisely for Fig. 3 we can write:

$$y_A(t) = y_I(t) + y_{IQ}(t)$$

or:

$$y_A(t) = \sum_k A_k h_I(t - kT) + \sum_k B_k h_{IQ}(t - kT)$$

and:

$$h_I(t) = g(t) \cos(\omega_c t) \otimes f(t) \cos(\omega_c t)$$

$$h_{IQ}(t) = g(t) \cos(\omega_c t) \otimes f(t) \sin(\omega_c t)$$
Since we are interested in having zero interference at sampling time \((kT)\), we look over \(h_{IQ}(t)\) in more detail:

\[
h_{IQ}(t) = \frac{1}{2} \int_{-\infty}^{\infty} g(\tau) \cos(\omega_c \tau) \cdot f(t-\tau) \sin \omega_c (t-\tau) d\tau \quad (12)
\]

or:

\[
h_{IQ}(t) = 0.5 \sin(\omega_c t) \int_{-\infty}^{\infty} g(\tau) f(t-\tau) d\tau + \frac{1}{2} \int_{-\infty}^{\infty} g(\tau) f(t-\tau) \sin \omega_c (t-2\tau) d\tau \quad (13)
\]

The second term in (13) is negligible because it is similar to filtering a high-frequency signal (modulated \(g(t)\) around \(2\omega_c\)) by a low-pass filter \(f(t)\). So we can write:

\[
h_{IQ}(t) \equiv \sin(\omega_c t) h(t) \quad (14)
\]

in which:

\[
h(t) = 0.5[g(t) \ast f(t)] \quad (15)
\]

Similarly it can be seen that:

\[
h_{I}(t) \equiv \cos(\omega_c t) h(t) \quad (16)
\]

Notice that in (14) \(h_{IQ}(t)\) is zero at \(t = 0\) which implies zero interference between corresponding symbols in \(I\) and \(Q\) paths. But we are interested in zero interference at other sampling times (i.e. \(t = kT, (k \neq 0)\)). One choice for this requirement is maintaining the *Nyquist Property* for \(h(t)\) which is the same condition as zero ISI in an equivalent PAM system. However, it should be noticed that this property here is more crucial because of its effect on interference between I and Q paths, whereas in PAM systems, adaptive equalizers and maximum likely-hood detectors can correct for ISI.

The second choice is:

\[
\sin(\omega_c kT) = 0 \quad k = \text{integer}
\]

or:

\[
2f_c < F_s = \text{integer}
\]

in which \(F_s = \frac{1}{T}\) is the baud-rate.

This pre-condition is practical, whenever we intend to use shaping filters \((g(t))\) with 100% excess bandwidth without any more serious restriction on them.

### 3. COMPARISON AND DIFFERENT ARCHITECTURES

Fig. 4 shows the CAP demodulator’s output \(y_A(t)\) and its components \(y_I(t)\) and \(y_Q(t)\), for a single complex symbol. In comparison to Fig. 2, we can see that \(y_I(t)\) has a narrower lobe which causes more sensitivity to the sampling instances. Also comparing \(y_{IQ}(t)\) we see that it is not near zero all along the time axis. Although \(y_{IQ}(t)\) is zero at sampling times (due to \(g(t)\) and \(f(t)\) being root raised cosine pulses), because of its slope near sampling points, the signal is more sensitive to sampling time error (or jitter).

Fig. 5 shows the demodulator’s output and the relevant sampled symbols for a sequence of symbols (4*4 Level), applied to both systems in Fig. 1 and Fig. 3 with the same condition of Fig. 2 and Fig. 4. As expected from (16), the CAP filter output (Fig. 5(b)) is not a baseband signal. Nevertheless, by maintaining some preconditions for CAP filters (see section 2.2), the transmitted symbols can be recovered by sampling at baud rate. However, the result is sensitive to sampling points and filters shape (as seen by the slope at sampling instances). Fig. 5(a) shows the output of a QAM demodulator which is like a PAM receiver’s output with the same input data as for Fig. 5(b). The output signal here is a baseband signal having lower slopes around sampling positions (hence less jitter sensitivity).

In Fig. 6 three architectures for CAP/QAM-type receivers have been shown. Most of the current CAP systems are implemented by digital I and Q filters, as what is shown in Fig. 6(a). In this way by having high orders filters the conditions in section B can be easily satisfied and a simple, efficient and flexible modem with fast start-up time will be obtained.[5]

Despite the above simplicity, for high-rate data transmission the design and implementation of the input A/D can be a major problem. For example for a baud rate around 150MHz, we need an A/D with a rate of about 450MHz. In addition, the resolution of the A/D needs to be higher because the symbol information are spread over the
whole input analog signal. Table (1) shows the effect of A/D resolution on the demodulator’s output error of a digital 16-CAP architecture.

In the case of higher rates, the architecture in Fig. 6(b) with analog CAP filters can be considered. Here, instead of a high-rate A/D, two A/D’s with 1/3 the sampling rate as well as lower resolution are used. However, the design and implementation of these analog CAP filters is a challenging problem as they must satisfy the preconditions discussed in section 2.2. As a result, the orders of these filters may be high. [6,7]

The last architecture is QAM with analog lowpass filters as shown in Fig. 6(c). This architecture has the benefits of less sensitivity to sampling jitter as well as easier analog filters and lower resolution A/Ds required. As we saw in section 2.1, a regular lowpass filter separates I and Q signals properly and so the filters are regular shaping filter for data transmission whose requirements can be relaxed through the use of adaptive equalization after the A/D converters [1]. However, the major problem here is adding analog multipliers which impose more circuit complexity and ensuring that the carrier phase and frequency are correct. Notice, simple square wave modulators can be used here, since the following lowpass filters will remove the higher order harmonics.

As a final comparison of A/D jitter effects in the demodulator’s output error, see Table 2. As expected, the QAM version output has less sensitivity to the sampling jitter. However, it should be noted that this comparison did not include jitter introduced at the analog modulators.

4. SUMMARY

In this paper we showed a comparison between CAP and QAM for near baseband data modulation. In addition, three different CAP-like architectures were discussed. For lower data-rate transmission, a mostly DSP CAP architecture has advantages because of simplicity and flexibility. At higher data-rates, two other architectures allow the use of lower-rate and lower-resolution A/D converters at the expense of some analog filtering and/or modulation. The best architecture choice is likely to be application dependent.

<table>
<thead>
<tr>
<th>No. of A/D bits</th>
<th>Error in dB</th>
</tr>
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<tbody>
<tr>
<td>10</td>
<td>-39.4</td>
</tr>
<tr>
<td>8</td>
<td>-30.12</td>
</tr>
<tr>
<td>6</td>
<td>-18.7</td>
</tr>
<tr>
<td>5</td>
<td>-16.6</td>
</tr>
<tr>
<td>4</td>
<td>-7.2</td>
</tr>
</tbody>
</table>

Table 1: Average Output Error for different A/D bit resolutions in digital 16-CAP

<table>
<thead>
<tr>
<th>Jitter interval (in psec)</th>
<th>CAP Digital</th>
<th>CAP Analog</th>
<th>QAM Analog</th>
</tr>
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<tbody>
<tr>
<td>0</td>
<td>-39.4</td>
<td>-38.9</td>
<td>-39.9</td>
</tr>
<tr>
<td>100</td>
<td>-36.1</td>
<td>-34.09</td>
<td>-39.3</td>
</tr>
<tr>
<td>200</td>
<td>-32.4</td>
<td>-31.4</td>
<td>-38.0</td>
</tr>
<tr>
<td>400</td>
<td>-27.6</td>
<td>-26.51</td>
<td>-33.9</td>
</tr>
<tr>
<td>1000</td>
<td>-19.8</td>
<td>-18.84</td>
<td>-27.05</td>
</tr>
<tr>
<td>2000</td>
<td>-13.4</td>
<td>-11.9</td>
<td>-21.12</td>
</tr>
</tbody>
</table>

Table 2: Average error of recovered symbols in dB, caused by sampling jitter, for different 16-CAP-like architectures

5. REFERENCES