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DIGITAL MATCHING OF RECEIVERS FOR AN ADAPTIVE-ANTENNA RECEIVING SYSTEM

by

K.H. Wu

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January 1997
Ottawa

Industry and Science Canada Industrie and Science Canada

The work was sponsored by the Department of National Defence, Research and Development Branch under Project No. 991-990-X21-718-998426156.

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Digital matching of receivers for an adaptive-antenna receiving system

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K.H. Wu
(Radio Communications Technologies Directorate)

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ABSTRACT

Mismatch of receivers in an adaptive-antenna receiving system can limit the interference-suppression performance. The precision of matching is critical in cases where the unwanted interference signals to be suppressed are much stronger than the desired signal. In this paper, a simple calibration technique for suppression of very strong interference signals is described.

RÉSUMÉ

La discordance des récepteurs dans un système d'antennes adaptives peut mener à une réduction de la capacité d'élimination de l'interférence. Une adaptation précise des récepteurs est cruciale dans les cas où les signaux d'interférence à éliminer sont beaucoup plus grands que le signal désiré. Dans cet article, une technique simple d'étalonnage est décrite qui mène à éliminer des signaux d'interférence très forts.

EXECUTIVE SUMMARY

In an adaptive-antenna receiving system, the amount of jamming suppression is limited by two possible sources, namely, how closely the receiver responses are matched, and imperfections in the particular signal processing algorithm used. The former source is attributable to the hardware and the latter to the signal processing software. The theme of this paper is how to alleviate the hardware limitation through receiver matching using a digital signal processing technique.

Suppression of unwanted interference signals must cover the bandwidths of the receivers in the antenna array. If parts of the frequency responses between receivers are different, then some of the unwanted interference signal can leak through regardless of how optimal the adjusted adaptive weights are.

Without receiver matching, the amount of mismatch could be in the order of 20 dB below the overall signal power, which is not sufficient for suppressing strong interference signals. In a high performance adaptive-antenna system, the mismatches should be in the order of 50 dB or more below the overall signal power. In the past, receiver matching was achieved by tweaking analog components in the receivers. This approach is usually laborious and expensive, and the level of matching precision achievable is rather limited.

In this paper, a simple technique for receiver matching by digital means is described. The technique was originally designed for the Programmable HF Adaptive-Antenna Receiving System (PHFAARS) developed for the Canadian Navy. It has been implemented in real-time software and installed on the PHFAARS. Although the technique was developed for matching HF receivers, it can be applied equally well to systems operating in other frequency bands.

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1. INTRODUCTION

In an adaptive-antenna receiving system, the frequency responses of all receivers are required to be precisely matched in both phase and amplitude in order to provide a high degree of interference suppression. The matching must cover the bandwidth of the receiving system.

In an ideal case of perfectly matched receivers in an antenna array, any undesired interfering signal can theoretically be cancelled completely, leaving only the desired communications signal and receiver thermal noise. In practice, analogue crystal IF filters commonly used in receivers have ripples in their amplitude and phase responses, which vary from receiver to receiver. In the past, matching of receiver responses was achieved by selecting a set of closely matched crystal filters or by tweaking the electrical characteristics of the filters. The process of matching analogue filters can be very costly, depending on the required precision of matching, and subject to degradation over time.

In this paper, a simple digital calibration technique is described. For matching a set of receivers, one receiver is chosen as a reference. A digital matching filter is designed for each of the rest of the receivers so that after filtering, their frequency responses closely resemble that of the reference receiver.

2. RECEIVER CALIBRATION REQUIREMENT

In order to completely cancel unwanted signals in a multichannel system, the frequency responses of the receivers must be identical over the bandwidths of the unwanted signals. In practice, no two receivers are the same and slight differences in the frequency responses will cause incomplete cancellation of the unwanted signals. In an adaptive cancellation scheme, a single complex weight is applied to each of the channels to adjust its amplitude and phase such that the difference across the signal bandwidth is minimized. It is apparent that if the frequency responses of the receivers are not identical, some of the unwanted signal will leak through regardless of how much the complex adaptive weights are optimized.

The *residual mismatch* leads to degraded performance. The measure of how far the unwanted signal can be suppressed is commonly known as the *null depth.* It is defined as the ratio of the residual unwanted signal after interference cancellation to the unwanted signal power before cancellation (see Equation (1) below). In this paper, both terms are used interchangeably, although in practice, the actual null depth could be limited by other factors.

The residual mismatch between two receivers, and hence the best null depth they can produce, can be calculated as the mean square difference between the frequency responses of the receivers over the bandwidths of the signal:

$$\left(\frac{\overline{|\varepsilon(t)|}^2}{P}\right)_2 \triangleq \frac{\int_{-\frac{B}{2}}^{\frac{B}{2}} |H_0(f) - H_1(f)|^2 df}{\int_{-\frac{B}{2}}^{\frac{B}{2}} |H_0(f)|^2 df} \tag{1}$$

where $H_0(f)$ and $H_1(f)$ are the complex frequency responses of the reference receiver and the receiver matched to the reference; B is the bandwidth of the signal of interest; P is the total power of the frequency response of the reference receiver to a unit power input signal; and $\overline{|\varepsilon(t)|}^2$ is the time-averaged signal-leakage power due to the difference between the two receivers. The subscript $_2$ indicates that there is only a single pair of receivers being considered.

The discrete form of Equation (1) is

$$\left(\frac{\overline{|\varepsilon(t)|}^2}{P}\right)_2 = \frac{\sum_k |H_0(k) - H_1(k)|^2}{\sum_k |H_0(k)|^2} \tag{2}$$

where k is a discrete frequency index and the summation is over the bandwidth of the signal, i.e. $K \cdot \Delta f = B$; K and Δf are, respectively, the total number of frequency samples and the frequency sample separation. For

$$\begin{aligned} H_o(k) &= A_o(k)e^{j\phi_o(k)} \\ H_1(k) &= A_1(k)e^{j\phi_1(k)} \end{aligned} \tag{3}$$

where A is the amplitude response, ϕ is the phase response in radians, and both are real variables,

$$\left(\frac{\overline{|\varepsilon(t)|}^2}{P}\right)_2 = \frac{\sum_k \left|A_o(k)e^{j\phi_o(k)} - A_1(k)e^{j\phi_1(k)}\right|^2}{\sum_k A_o^2(k)}$$

$$= \frac{\sum_k \left| e^{j\phi_1(k)} \left[A_o(k) e^{j[\phi_o(k)-\phi_1(k)]} - A_1(k) \right] \right|^2}{\sum_k A_o^2(k)}$$

$$\cong \frac{\sum_k \left| e^{j\phi_1(k)} \left\{ A_o(k) \left[1 + j(\phi_o(k) - \phi_1(k)) \right] - A_1(k) \right\} \right|^2}{\sum_k A_o^2(k)}$$

for | $\phi_0(k)$ - $\phi_1(k)$ | < 0.1 radian,

$$= \frac{\sum_k \left[A_o(k) - A_1(k) \right]^2 + \sum_k A_o^2(k) \left[\phi_o(k) - \phi_1(k) \right]^2}{\sum_k A_o^2(k)}$$

$$= \frac{\sum_k \left[A_o(k) - A_1(k) \right]^2}{\sum_k A_o^2(k)} + \frac{\sum_k A_o^2(k) \left[\phi_o(k) - \phi_1(k) \right]^2}{\sum_k A_o^2(k)}$$

If $A_0(k)$ is more or less constant with respect to k, as in most receivers, then

$$\left(\frac{\overline{|\varepsilon(t)|^2}}{P} \right)_2 \cong \frac{\sum_k \left[A_o(k) - A_1(k) \right]^2}{\sum_k A_o^2(k)} + \frac{\sum_k \left[\phi_o(k) - \phi_1(k) \right]^2}{N}$$

$$= \frac{\sum_k \Delta A^2(k)}{\sum_k A_o^2(k)} + \frac{1}{N} \sum_k \Delta\phi^2(k) \tag{4}$$

where $\Delta A(k)$ = | $A_0(k)$ - $A_1(k)$ |, $\Delta\phi(k)$ = | $\phi_0(k)$ - $\phi_1(k)$ |, and N is the number of frequency samples used in the calibration process. Equation (4) provides an approximate but convenient method to estimate null depth based on the sum of the separate amplitude and phase mismatch quantities.

A contour plot of null depth, as originally defined in Equation (1) and estimated in Equation (4) for the two-receiver case, is shown in Figure 1. In the figure, the vertical axis is the root mean square phase difference in degrees and the horizontal axis is the root

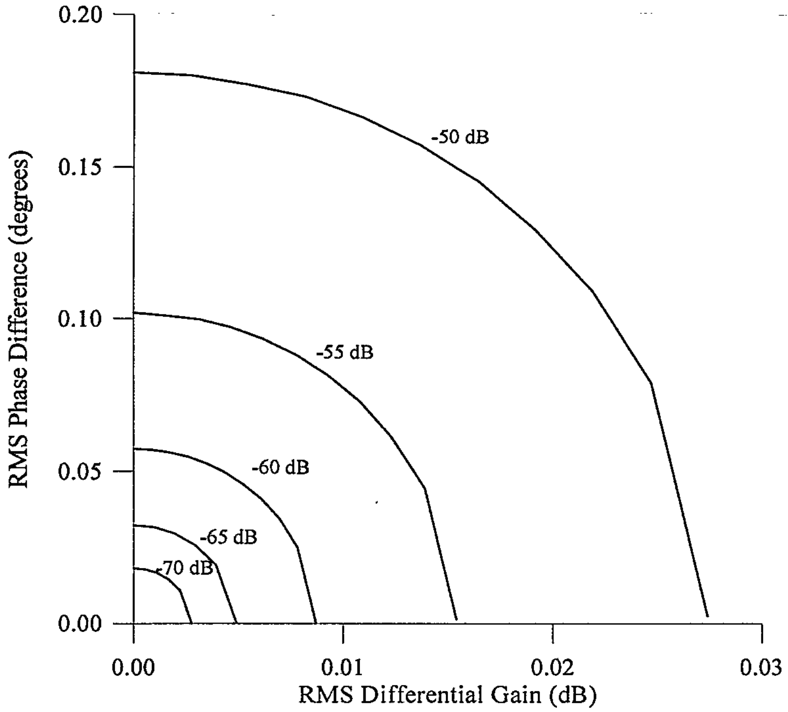


Figure 1: Constant Null-Depth Contours for a Single Pair of Receivers.

mean square differential gain in dB. The rms differential gain in dB on the horizontal axis is defined as $20\log_{10}\left(\sqrt{\frac{\sum_k \Delta A^2(k)}{\sum_k A_0^2(k)}}+1\right)$, and rms phase difference on the vertical axis is defined as $\frac{180}{\pi}\sqrt{\frac{1}{N}\sum_k \Delta\phi^2(k)}$ in degrees. The rms differential gain is defined in this way primarily for ease of comparison, since most filter or receiver frequency responses are given in dB. It should not be confused with the calculation of the residual mismatch as defined in Equation (1), where the *difference* in complex amplitude rather than ratio is defined.

The above derivation is based on one pair of receivers. In a system consisting of L receivers, the amount of residual mismatch at the summed output of an array system will increase with the number of receivers. More precisely, if we define $\overline{|\varepsilon_w(t)|^2}$ as the residual mismatch of the worst selection of a receiver pair from the L receivers, then an upper bound for a system incorporating all L receivers is

$$\left(\frac{\overline{|\varepsilon(t)|^2}}{P}\right)_L \leq \frac{L}{2}\left(\frac{\overline{|\varepsilon_w(t)|^2}}{P}\right)_2 \tag{5}$$

Therefore, if we use four receivers that are equally mismatched, the residual mismatch at the output will increase by 3 dB in power.

3. CALIBRATION SCHEME

The proposed calibration scheme uses a sequence of stepped tones scanning across the receiver bandwidth. A block diagram for the calibration signal collection is shown in Figure 2, where a four-channel receiving system is shown. The calibration signal consists of a sequence of N equally spaced frequency tones generated by either a direct digital synthesizer (DDS) or any other suitable synthesizer. The calibration signal is split equally and fed into the RF inputs of the receivers. Only one tone is present at a time. The frequency of the DDS is controlled by a digital controller or computer. During calibration, the RF inputs of the receivers are switched from the antenna outputs to the DDS outputs, i.e. all receivers receive the same calibration signal. At each frequency step, a number of time samples, J, are collected for data averaging and evaluation purposes. If N frequency tones are used to cover the receiver bandwidth, there are $J \times N$ data collected for each channel.

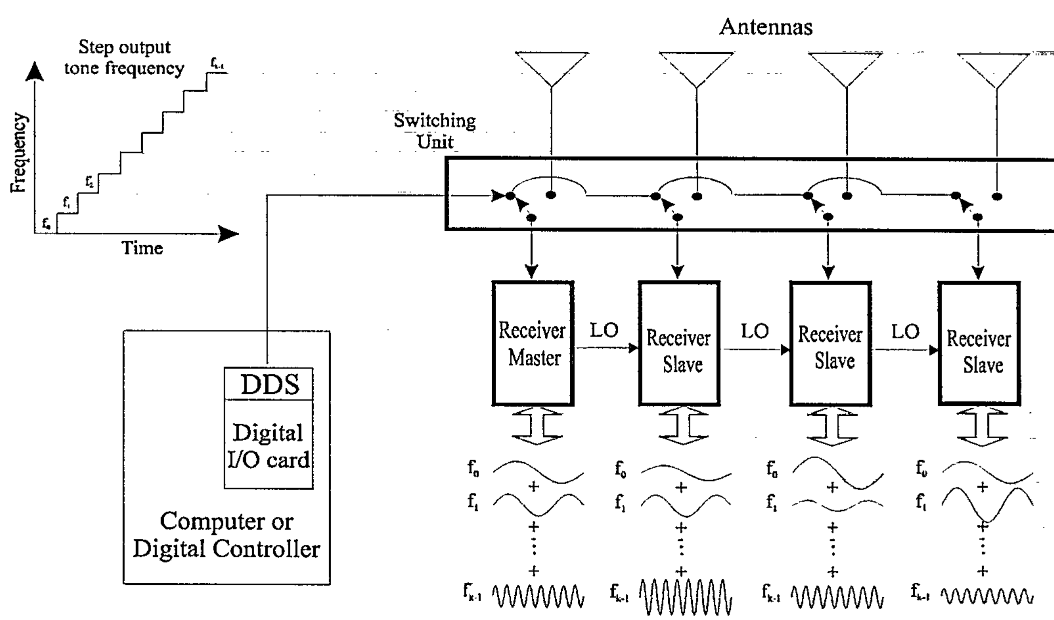


Figure 2: Calibration Signal Collection.

4. CALCULATION OF MATCHING FILTER COEFFICIENTS

The calibration filters are of finite-impulse response (FIR) type. As in most antenna-array applications, the received data are complex, i.e. with in-phase and quadrature components. A direct approach for generating the calibration filter coefficients, $c(i)$, would be to take an inverse Fast Fourier transform (FFT) of the ratio of the frequency responses of the reference receiver to the receiver to be matched, viz.

$$Y(f) = \begin{cases} \dfrac{H_0(f)}{H_1(f)}, & \text{within the receiver bandwidth} \\ \text{undefined,} & \text{outside the receiver bandwidth} \end{cases} \tag{6}$$

$$c(i) = \Im^{-1}[Y(f)] \tag{7}$$

where $\Im^{-1}[\bullet]$ denotes the inverse Fourier transform operation, and $H_0(f)$ and $H_1(f)$ are the frequency responses of receivers 0 and 1 respectively. The values of $Y(f)$ outside the receiver bandwidth are undefined since the values of $H_0(f)$ and $H_1(f)$ are close to zero there. A convenient but not optimal way is to pad the undefined region with zeros. The main theme of this paper is how to fill the undefined region so that $c(i)$ is represented by as few significant samples as possible, since these samples are used as the matching-filter coefficients. Receiver 0 is picked as a reference and receiver 1 is to be calibrated so that its calibrated frequency response closely resembles that of receiver 0. If J time samples are collected for each tone frequency as described above, the ratios, H_0/H_1, for the samples can be averaged to yield a better signal-to-noise ratio before the inverse Fourier transform operation. The output, $c(i)$, are the required matching filter coefficients and i is the coefficient index. When the coefficients are applied to the calibration data input to receiver 1, the filtered data are matched to those of receiver 0.

If the undefined region is padded with zeros, $c(i)$ would contain many significant sidelobes because of the abrupt discontinuities at both ends of the undefined region. The sidelobes of $| c(i) |$ decay slowly with index i. Hence the number of coefficients required to meet a specific matching criterion would be large. This means that more processing power is required in a real-time environment. This implies that the number of significant coefficients (i.e. the coefficients that are not small enough to be ignored) is large.

The values of $Y(f)$ outside the signal bandwidth are undefined and therefore they can be set to arbitrary values that give insignificant sidelobes in $c(i)$. This can be achieved by filling the undefined region in the frequency domain with a smooth continuous curve joining both ends of $Y(f)$ at the edges of the receiver bandwidth. This is equivalent to an interpolation in which the exact values are not relevant as long as the overall function of

$Y(f)$ is smooth and without discontinuities. The interpolation can be achieved by fitting a third-order polynomial curve to both ends of the signal bandwidth:

$$Y(k) = b_o + b_1 k + b_2 k^2 + b_3 k^3, \qquad \frac{N+1}{2} \le k \le K - \frac{N+1}{2} \tag{8}$$

where K is the length of the inverse FFT and the negative portion of $Y(k)$ is assumed to be in the upper portion on the k axis, $K - \frac{N-1}{2} \le k < K$, and N is an odd integer. To perform the interpolation, two values of $Y(f)$ taken from each edge are used to calculate the polynomial coefficients:

lower edge samples at $k_0 = \frac{N-3}{2}$ and $k_1 = \frac{N-1}{2}$,

upper edge samples at $k_2 = K - \frac{N-1}{2}$ and $k_3 = K - \frac{N-3}{2}$.

The polynomial coefficients, b_0, b_1, b_3, b_4 are then given by

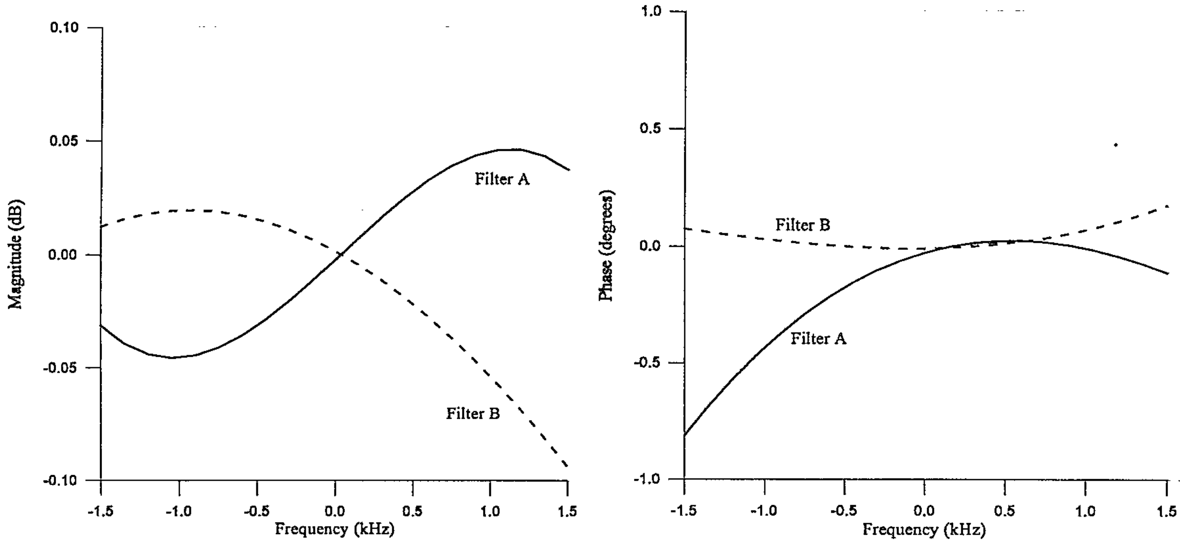
$$\begin{bmatrix} b_0 \\ b_1 \\ b_2 \\ b_3 \end{bmatrix} = \begin{bmatrix} 1 & k_o & k_o^2 & k_o^3 \\ 1 & k_1 & k_1^2 & k_1^3 \\ 1 & k_2 & k_2^2 & k_2^3 \\ 1 & k_3 & k_3^2 & k_3^3 \end{bmatrix}^{-1} \begin{bmatrix} Y(k_o) \\ Y(k_1) \\ Y(k_2) \\ Y(k_3) \end{bmatrix} \tag{9}$$

5. A DESIGN EXAMPLE

An example is given here to illustrate the above design procedure. A major source of mismatch for common receivers originates from the IF crystal filter in the first stage of frequency conversion. The frequency responses of two typical crystal filters used in HF receivers are used in this example: Piezo Technology Inc., Model 8069,.at 40.455 MHz. In fact, two filters with the largest differences in amplitude and phase were picked from a pool of about 30 filters, so this example represents a case of bad mismatch.

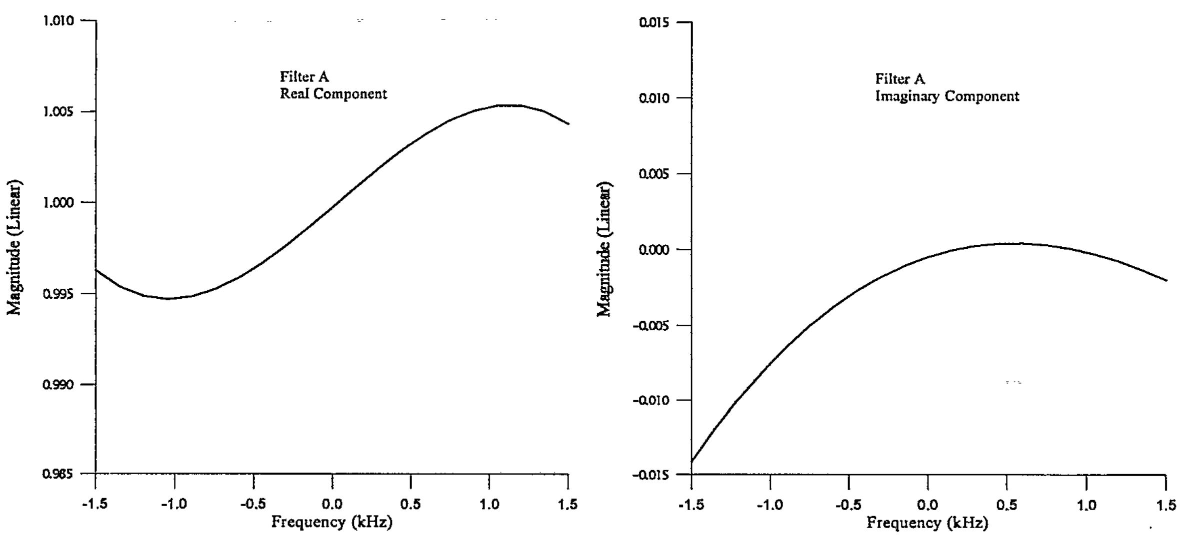
The magnitude and phase responses of the two filters, designated as Filters A and B, are shown in Figures 3(a) and 3(b) respectively, where magnitude and phase are plotted versus frequency. The bandwidth of interest is 3 kHz, which occupies the centre portion of the crystal filter bandwidth of 20 kHz. Without matching these two receivers, the null depth would be approximately -43 dB.

To facilitate computation, the magnitude and phase responses of Filters A and B are first converted into real and imaginary components, i.e. complex numbers. The resulting complex responses for Filters A and B are plotted in Figures 4 and 5, respectively.



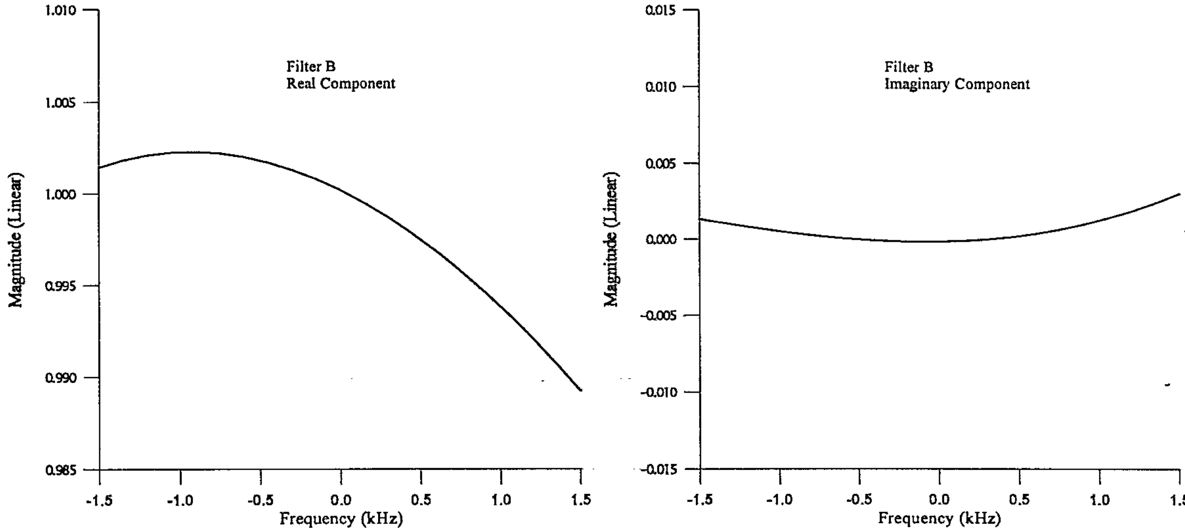
(a) (b)

Figure 3: Magnitude and Phase Responses of Two Badly Mismatched Crystal Filters, A and B.



(a) (b)

Figure 4: Real and Imaginary Components of the Frequency Response of Filter A.

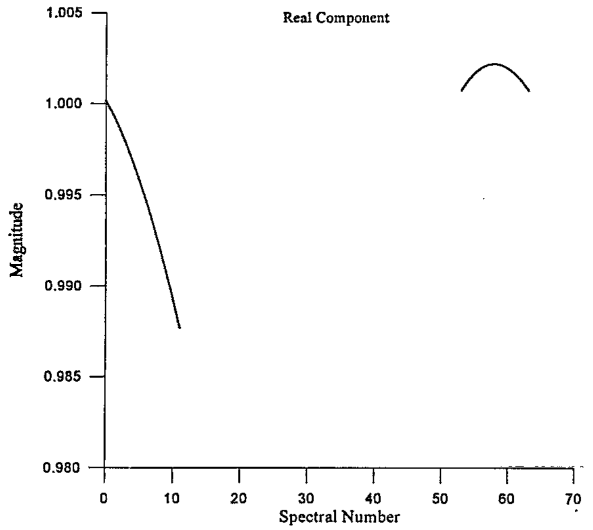


(a) (b)

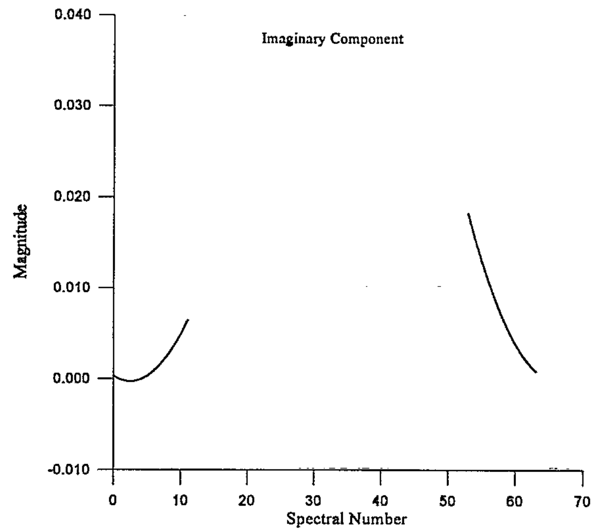
Figure 5: Real and Imaginary Components of the Frequency Response of Filter B.

Assuming a time sampling rate of 9,600 complex samples per second and an FFT length (K) of 64, the test tones are equally spaced with a frequency step of 9,600/64, or 150 Hz. The sampled frequency response of Filter A is then divided by that of Filter B, as given in Equation (6). The result is plotted in Figure 6. A total of 23 test tones (N) are used to sample the spectra of the filters to cover a bandwidth of slightly larger that the required 3-kHz signal bandwidth. In the figure, the upper half of the frequency axis actually represents the negative frequency axis. This is to facilitate the curve fitting and the FFT operations described below. The centre portion is outside the bandwidth of the signal and is left empty. As described above, if the empty area is padded with zeros and an FFT operation taken, the abrupt discontinuities at both edges of the zero-padded region would produce significant sidelobes in the time domain. These sidelobes, however, must be included in the time-domain filter coefficients to reduce the amount of residual mismatch after filtering; this in turn increases the number of taps in the filter.

In a real-time application, it is important to have as few filter taps as possible while maintaining the amount of residual mismatch to an acceptable level. To reduce the sidelobe level in the time domain, the corresponding frequency response should have no abrupt discontinuities and must vary smoothly with respect to frequency. To satisfy these requirements, the "empty" area in the frequency response can be filled with a smooth curve by means of curve fitting. Since no signal exists in the empty area, the exact shape of this curve is irrelevant as long as the overall frequency response is smooth. Incidentally, the curve-fitted frequency response now becomes that of an all-pass filter.



(a)



(b)

Figure 6: Sampled Frequency Response of Filter A Divided by Filter B (i.e. Y(f) in Equation (7)). The negative portion is moved to the upper part of the (cyclic) frequency response.

The curve fitting is accomplished by using a third-order polynomial equation connecting both ends of the frequency response ratio of the two filters, as described by Equations (8) and (9). Since the frequency response is complex, the curve fitting must be done on both the real and imaginary parts. For a third-order curve fitting, four samples are needed to calculate the polynomial coefficients. The obvious choice of samples is the two samples at each end of the frequency response. The samples used for computing the b coefficients in Equation (9) are $k_0 = 10$, $k_1 = 11$, $k_2 = 53$, $k_3 = 54$. The curve-fitted frequency response of Figure 6 is shown in Figure 7. The new frequency response is now two continuous smooth curves in both complex components. This helps to reduce the sidelobe level in the time domain significantly. Hence, for a given maximum sidelobe level, the number of filter coefficients can be reduced.

The upper end of the fitted curve (sample #52) does not join the original curve (sample #53) exactly as it should. The slight glitch is due to precision error in the computation of the third-order polynomial curve fitting. Despite this error, the overall residual mismatch error is relatively small in this example.

The amount of the resulting residual magnitude and phase mismatches versus frequency is plotted in Figure 8. Shown in the figure are also the mismatch errors before calibration for comparison. The vertical axis in Figure 8(a) is in absolute dB units and the tick labels are in log scale (i.e. the magnitude mismatch is shown in a double log scale). In Figure 8(b), the vertical axis is the phase mismatch in degrees. Mismatches for two calibration filters with 9 and 21 coefficients are shown in the figure. The mean square residual mismatch error versus number of filter coefficients ranging from 7 to 25 is plotted in Figure 9.

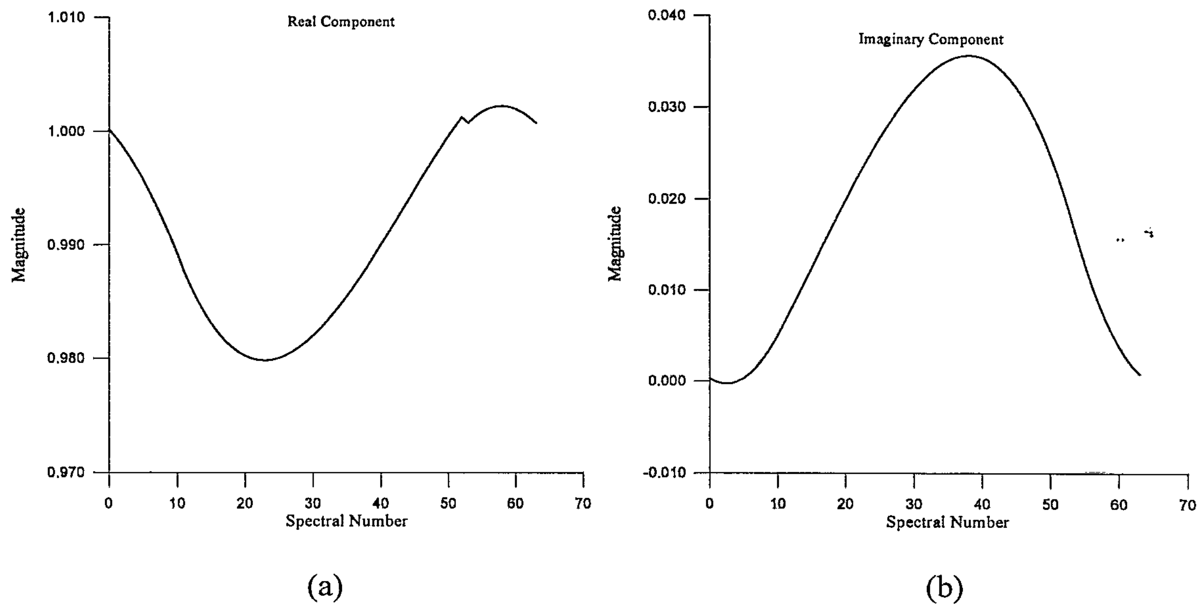


Figure 7: Curve-Fitted Frequency Response of that Shown in Figure 6.

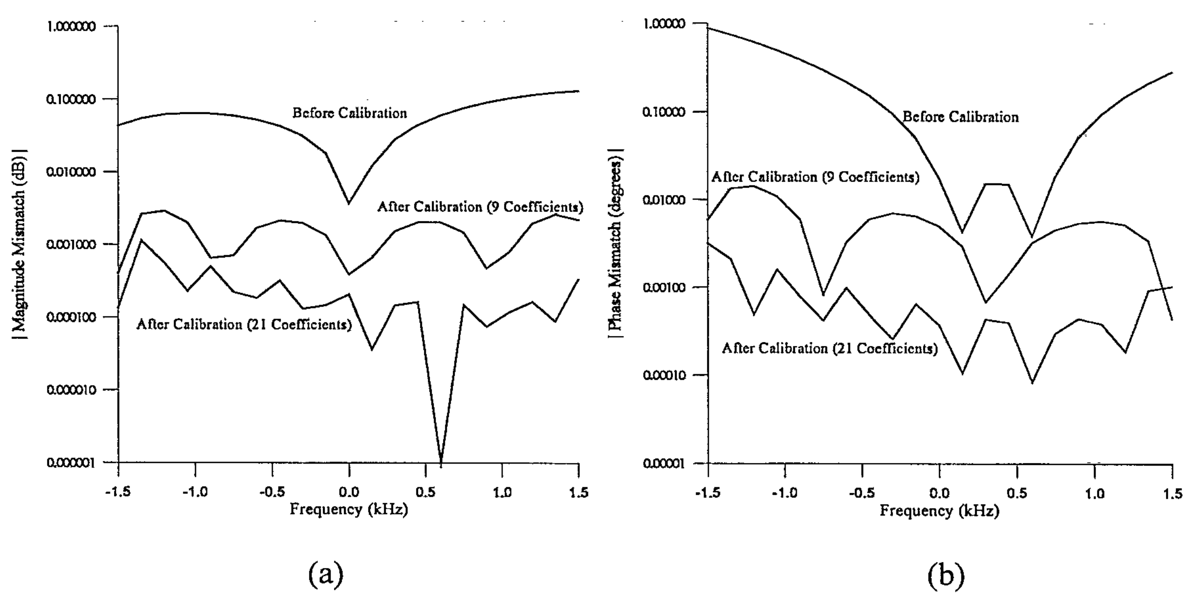


Figure 8: Residual Mismatches in Magnitude and Phase.

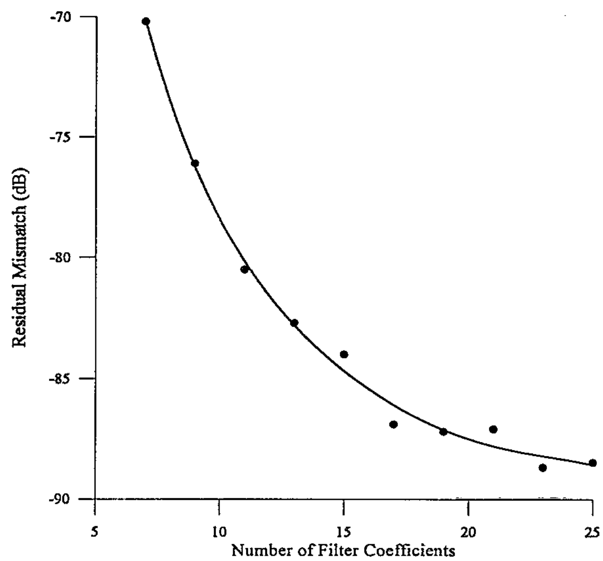


Figure 9: Mean Square Residual Mismatch Error Versus Number of Filter Coefficients.

To examine how the mean square residual mismatch increases with increasing rms phase difference, the latter parameter in this example was amplified by a factor of up to ten. The results for 9-coefficient and 21-coefficient filters, as well as for no calibration, are shown in Figure 10. Also shown in the figure is the result for a 21-coefficient filter with zero padding in the no-signal region in the frequency response, i.e. without curve fitting. These curves show that receiver calibration is mandatory if a null depth of better than -50 dB is required. A nine-tap filter can provide a null depth in the order of -60 to -70 dB, which is sufficient for most high-performance adaptive-antenna receiving systems.

6. CONCLUSIONS

The residual mismatch due to the difference in the frequency responses of receivers limits the interference suppression performance of adaptive-antenna receiving systems. A simple method of receiver calibration was designed to reduce the mismatch and thereby enhance the performance in the presence of strong interference signals. A design example was given to illustrate the design procedure. It was shown that a third-order polynomial curve-fitting approach can reduce the number of matched-filter coefficients while at the same time reducing the amount of residual mismatch after receiver calibration.

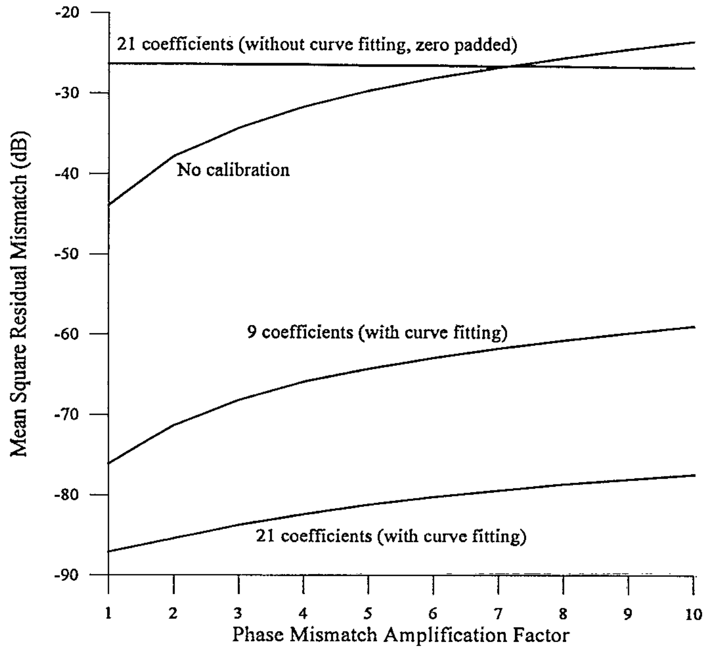


Figure 10: Mean Square Residual Mismatch Versus Phase Mismatch Amplitude.

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MISMATCH OF RECEIVERS IN AN ADAPTIVE-ANTENNA RECEIVING SYSTEM CAN LIMIT THE INTERFERENCE-SUPPRESSION PERFORMANCE. THE PRECISION OF MATCHING IS CRITICAL IN CASES WHERE THE UNWANTED INTERFERENCE SIGNALS TO BE SUPPRESSED ARE MUCH STRONGER THAN THE DESIRED SIGNAL. IN THIS PAPER, A SIMPLE CALIBRATION TECHNIQUE FOR SUPPRESSION OF VERY STRONG INTERFERENCE SIGNALS IS DESCRIBED.

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