Winter 12-1-1988

Broadbanding Techniques for Microstrip Patch Antennas - a Review

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SCIENTIFIC REPORT NO. 98

BROADBANDING TECHNIQUES FOR MICROSTRIP PATCH ANTENNAS - A REVIEW

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December 1988

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This report is based on a presentation of the same topic given at the 1988 Antenna Applications Symposium, Robert Allerton Park, University of Illinois, Sept. 21-23, 1988.

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Contents

	Abstract	1
1.	INTRODUCTION	1
	Bandwidth Definitions	2
2.	BANDWIDTH OF SINGLE PATCH ANTENNAS	3
	Non-rectangular Shaped Patches	5
3.	SINGLE SUBSTRATE MULTIPLE-RESONATOR CONFIGURATIONS	5
4.	MULTIPLE-RESONATOR CONFIGURATIONS WITH PATCHES STACKED VERTICALLY	8
5.	IMPEDANCE MATCHING NETWORKS FOR BROAD-BAND MICROSTRIP ANTENNAS	11
	Electromagnetically Coupled Microstrip Patches	14
6.	BANDWIDTH CONSIDERATIONS FOR CIRCULARLY POLARIZED PATCHES	15
7.	BROAD-BAND MICROSTRIP ARRAY CONFIGURATIONS	17
8.	CONCLUDING REMARKS	18
	References	20
	Figures 1 through 12	25-30

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Abstract

The need for development of wideband microstrip patch antennas has been very well recognized. This article is a review of various techniques attempted for increasing the bandwidth of microstrip patches. These include multiple resonator configurations (with patches located on the same plane as well as multiple-layered substrate geometries) and wideband impedance matching network approach. Various results reported in the literature are discussed and areas for continued research efforts are identified.

1. INTRODUCTION

Narrow bandwidth available from printed microstrip patches has been recognized as one of the most significant factors limiting the widespread applications of this class of antennas. The operating bandwidth of a single linearly polarized patch antenna is limited by its input VSWR (or reflection loss) and is inversely proportional to the Q-factor of the patch resonator. Typical values for the fractional bandwidth are 2 to 4%. Over the last decade, several attempts have been made for improving the bandwidth of microstrip patches. This article is a review of various techniques employed for broadband design of microstrip antenna elements and arrays and attempts to present a comparison of the results reported.

Various techniques used to increase the bandwidth of microstrip patch antennas may be classified as follows: (i) Decreasing the Q-factor of the patch by increasing the substrate height and lowering the dielectric constant. (ii) Use of multiple resonators located in one plane. (iii) Use of multilayer configurations with multiple resonators stacked vertically. (iv) Use of impedance matching networks.

These techniques are discussed in Sections 2 through 5. Special considerations for the bandwidth of circularly polarized patches are discussed in Section 6. Finally, bandwidth considerations for arrays of microstrip patches are contained in Section 7. Comparison of various approaches and concluding remarks are contained in the last section.

Bandwidth Definitions

Since the bandwidth (BW) of linearly polarized single microstrip patches is usually limited by the reflection loss at the input terminal, the commonly used definition for BW of the microstrip patches is *the frequency range over which the input VSWR is less than certain specified value s*. Usually s is taken equal to 2 but certain specifications choose s = 1.5. Thus one has to be careful while comparing bandwidths quoted in different reports.

A concept of *radiation bandwidth* for microstrip antennas has been introduced in [1]. The radiation BW is defined as the frequency range over which the radiated power is within 3 dB of the incident power and the radiation pattern is essentially the same. The second part of this definition can be made more specific by quantifying the allowed variations in the radiation pattern (for example in beam direction, beam width, or the sidelobe level). This definition is more comprehensive as the radiated power is reduced because of the input reflection as well as by other factors such as changes in voltage distribution along the edges of the patch.

2. BANDWIDTH OF SINGLE PATCH ANTENNAS

Input impedance and the impedance bandwidth of a microstrip patch antenna may be computed using its transmission line model or the cavity model. If we represent the resonant patch in terms of a parallel L-C-G network, its input admittance Y_{in} can be written as

$$Y_{in} = G(1 + 2jQ\,\Delta f/f_0) \tag{1}$$

where $Q = \omega C/G$ and Δf is the deviation from the center frequency f_0. Bandwidth BW defined for input VSWR to be better than s is given [2] by

$$BW = \frac{s-1}{Q\sqrt{s}} \tag{2}$$

For s=2, and substituting for Q in terms of energy stored and power radiated we can write

$$BW \approx \frac{\sqrt{2}\,h\,G_e}{\pi c \sqrt{\varepsilon_{re}}\;\varepsilon_0 b_e} \tag{3}$$

where G_e is the edge conductance, h is the substrate thickness, b_e is the effective width and ε_{re} the effective dielectric constant of the equivalent microstrip line. From (3), we note that the BW increases linearly with increase of h. Also, BW increases when ε_r (and hence ε_{re}) is reduced. Recognizing that the edge conductance for a patch width b is given by [3],

$$G_e = \frac{1}{120}\frac{b}{\lambda_0} \quad \text{for} \quad b/\lambda_0 \gg 1 \tag{4}$$

and

$$G_e = \frac{1}{90}\left(b/\lambda_0\right)^2 \quad \text{for} \quad b/\lambda_0 \ll 1$$

we find the fractional bandwidth BW increases with the operating frequency f_0.

Typical variation of BW with h/λ is depicted in [4] whereas variation with frequency is shown in [2]. Computations agree with the measured values. As pointed out in [3], bandwidths up to 15% can be obtained at the expense of an increase in the antenna height to about 0.1 wavelengths. However, such thick substrates are not desirable or practical because of the following reasons: (i) For probe-fed patches, an increase in the substrate thickness causes an increase in the probe inductance which, in turn, creates input matching problems. (ii) For microstrip-fed patches, increased substrate thickness causes an increase junction reactance, which creates spurious radiation as well as the input match problems. (iii) Thick substrates are not well suited for microstrip circuitry used for signal distribution and phase shifting networks fabricated on the same substrate. (iv) Thick substrates make it mechanically difficult to have antenna arrays conformal to curved surfaces (of aircraft, space craft, missiles, etc.). (v) Many of the analysis and design techniques used (cavity model, etc.) become inaccurate for thick substrates.

Decrease in the BW of microstrip antenna by decreasing the substrate thickness h, or decreasing the width b, or by decreasing the length a (by increasing ε_r) is consistent with the general relationship between the Q and volume of small antennas [5].

A useful formula for bandwidth of microstrip patches is given by Munson [6] as

$$BW = 4f^2 \left(\frac{t}{1/32}\right) \tag{5}$$

where BW is the bandwidth in MHz (for VSWR < 2), f is the operating frequency in GHz, and t is the thickness in inches.

Non-rectangular Shaped Patches

Bandwidths available from circular microstrip patches [7] are of the same order as those available from rectangular or square shaped antennas. The two other geometries which have been shown to yield wider bandwidths are: circular ring [8] and square ring [9] microstrip antennas. As pointed out in [8], TM_{11} mode of the annular ring is a poor radiator but TM_{12} mode of the ring yields a bandwidth about two times larger than the bandwidth of a circular disk. The bandwidths for the rectangular patch and for the rectangular ring patch are compared in [9], and in a typical case ($\varepsilon_r = 2.50$, h = 0.159 cm, and f = 1080 MHz), the ring has a BW value 1.63 times that of the rectangular patch. Qualitative reasons for improved bandwidth are the same in two cases. For ring geometries, energy storage underneath the patch is smaller but the power radiated from the radiating edges is about the same.

3. SINGLE-SUBSTRATE MULTIPLE-RESONATOR CONFIGURATIONS

Use of multiple resonant patches located in the same plane (on the top surface of a single substrate) has been explored by several investigators [10-16]. The key idea in these configurations is that the stagger tuned coupled resonators can yield wider frequency response in the same way as obtained in the case of multi-stage tuned amplifier circuits.

In [10], two narrow conducting strips are placed close (~2.5 to 3.0 times the substrate thickness) to the the non-radiating edges of a rectangular patch (Fig. 1). Bandwidth of 5% (VSWR < 2, f_0 = 350 MHz, on 1.27 cm thick silicone fiberglass substrate) was obtained experimentally. Coupling of parasitic patches to the radiating edges of a rectangular patch has been described in [11]. In this case, the parasitic patches are $\lambda/4$ long (shorted at the far ends, as shown in Fig. 2) and therefore the locations of the radiating apertures do not change. A bandwidth 2.12 times that of a single open patch (f_0 =1.275 MHz, ε_r = 2.5, h = 3.18 mm, VSWR < 2) has been reported. The characteristics of this configuration have been explained in terms of an antiphase mode of a pair of coupled resonators and it is shown that the bandwidth improvement is independent of the coupling capacitance.

Multiport network analyses of multiresonator configurations (Fig. 3 a,b and c) with additional patches gap coupled to the radiating edges, to the non-radiating edges, or to all the four edges of a probe-fed central patch are reported in [12] and [13]. Coupling gaps are modeled by multiport capacitive networks and two-dimensional analysis (segmentation method) is used to evaluate input impedance as well as the voltage distributions around the edges. Bandwidth (VSWR < 2) of the five-resonator configuration (Fig. 3c) has been evaluated theoretically and experimentally to be 6.7 times that of a single patch on the same substrate (ε_r = 2.55, h = 0.318 cm, and f_0 = 3.16 GHz, BW = 25.8%). Alternative multiport configurations wherein the additional resonators are not gap-coupled but connected to the central patch through small sections of microstrip lines (as shown in Fig. 4) have been discussed in [14]. Again multiport network modeling and planar analysis is employed and experimental verifications are reported. For the 5-patch configuration (Fig. 4(c)), a bandwidth improvement of 7.36 times over that for a single patch (VSWR < 2, ε_r = 2.55, h = 0.318 cm, f_0 = 3.18 GHz, BW = 24%) have been reported.

Another interesting multi-patch antenna configuration is described in [15]. This configuration consists of 4 triangular patches as shown in Fig. 5. The central patch A is probe fed, the lower patch B is gap coupled, and the other two patches C and D are coupled by short sections of microstrip lines. The design was optimized by experimental iterations. A bandwidth (VSWR < 2) of 11.3% (ε_r = 2.55, h = 1.6 mm, f_0 = 3.19 GHz) which is 5.4 times the bandwidth of a single rectangular patch antenna has been reported. Over this bandwidth, the 3 dB beamwidth varies from 55° to 72° in the E-plane and 40° to 70° in the H-plane. The cross-pol level is better than 14 dB in the H-plane and 12 dB in the E-plane. Analysis and design of this type of antenna may also be carried out by the multiport network modelling approach discussed for multiresonator rectangular patches in [12-14]. Such antennas may find use in applications where a wide bandwidth is required but some variations in radiation characteristics over the bandwidth range may be tolerated.

There are two problems associated with the use of the multiresonator configurations described above: (i) larger area requirement and consequent difficulty in using these configurations as array elements, and (ii) variations of the radiation pattern over the impedance bandwidth of the configuration.

A modification of the multiresonator patches (for avoiding the above-mentioned two problems) is described in [16]. This configuration (shown in Fig. 6) is a modification of the non-radiating-edges coupled resonator configuration (shown in Fig. 3b). Six narrow patches are coupled through capacitive gaps. Their lengths are equal but widths are different. The configuration was designed by experimental iterations. Measured bandwidth (VSWR < 2) equal to 10 times that of a single patch of the same overall area has been reported (ε_r = 2.2, h = 0.8 mm, f_0 ~ 875 MHz, BW ~ 6%). Since the size of this multiresonator configuration is compact, the radiation pattern is consistently good over the bandwidth. Measured radiation patterns at 854 MHz and 897 MHz support this contention. Information regarding variation of gain over the bandwidth is not available in [16]. Analysis and design procedures for this type of antenna have not been reported so far. Of course, the multiport network modeling approach used for 3-patch configuration of Fig. 3b can be extended to the six-patch configuration also. An alternative approach is to model the multiple-resonator patch as a section of a multiple coupled microstrip line, and to construct a coupled-line model (similar to the transmission line model for a single patch). This concept of coupled-line model as applicable to a two-resonator configuration is discussed in [1]. Certainly, multiple-resonator configuration with narrow patches coupled along non-radiating edges, is an antenna configuration for which further detailed investigations are recommended.

4. MULTIPLE-RESONATOR CONFIGURATIONS WITH PATCHES STACKED VERTICALLY

In this approach, two or more layers of dielectric substrates are used. Resonant patches are located on the top of each of the substrate layers and are stacked vertically. Two-layer configurations are most common, although three dielectric layers(and three patches) have also been used. The two patches may be identical (same size) or of slightly different sizes (and hence resonant at slightly different frequencies). When the patches have unequal dimensions, two different arrangements are possible. Smaller patches can be located on the top layer (Fig. 7a) or the larger patch can be placed on the top (Fig. 7b). Both of these arrangements have been used. When the smaller patch is on the top, edges of both the smaller and the larger patches become the sources of radiation with the effective aperture shifting from the bigger patch to the smaller patch as the frequency of operation is increased. On the other hand, when the larger patch is on the top (and the smaller patch below), the boundary of the upper patch constitutes the radiating aperture. The lower patch helps in the broadband excitation of the upper patch, and is termed as the feeder patch. Of course, two patches can have identical dimensions in which case the distinction between the feeder and the radiator patches disappears and the two functions are merged together. In most of these two-patch configurations, the lower patch is fed via a probe or a microstrip line. Two different ways of excitations of the upper patch are available. We can use a vertical probe passing through a hole in the lower patch and making contact to the upper patch. Alternatively capacitive coupling between the patches is used to excite the upper patch.

Double layer two-patch configurations were initially used for dual frequency microstrip antennas [10,17]. Wide bandwidth application was reported by Hall et al in [18] wherein two-layer and three-layer configurations are described. In these designs, the lower substrate layer is alumina (ε_r = 9.8, h = 0.625 mm) where the feed line and the associated circuitry is fabricated. The second and third layers are polyguide (ε_r = 2.32, h = 1.59 mm) where the resonant radiating patches are fabricated (see Fig. 8 a and b). Bandwidths of 13% and 18% (return loss > 10 dB, f_0 ~ 10.6 GHz) were measured for two-layer and three-layer configurations respectively. For comparison, bandwidth of a single layer patch on alumina substrate is only 1.1%, while that on polyguide is about 6.6%. This technique of capacitively-coupled stick-on antennas is well suited for certain applications where the antenna may be regarded as an additional component in the transmitter/receiver circuits.

There have been several reports of two layer antennas with an air-gap in between the two layers. In the configuration described in [19], the lower patch has a smaller size and acts as a feeder resonator. Out of the various configurations presented, S-band circular disc configuration has a bandwidth of 15% (VSWR < 2.0, air-gap thickness not specified). For the configuration reported in [20], a low ε_r dielectric foam material is used to separate the two patches. For a spacing of 0.572 cm, a 17.3% BW (VSWR < 1.92, 3.85-4.58 GHz) is reported. The lower patch is smaller in size and is fed by a probe or a microstrip line. The element has been used in 4x4 array of circularly polarized patches (two feed points on each patch). An investigation of the two layer antenna performance as a function of the air-gap spacing is discussed in [21].

An experimental study of vertically stacked triangular patches is reported in [22]. Both dual frequency and broadband operation are described. For two identical equilateral triangular patches fabricated on dielectric layers (with $\varepsilon_r = 2.55$ and thickness 1.6 mm) separated by a foam thickness of 5.0 mm, a bandwidth of 17.46% (VSWR < 2, $f_0 = 3.61$ GHz) has been reported. For comparison, the bandwidth of the same two-patch configuration with zero foam thickness is only 2.5%.

All of the vertically-stacked multiple resonator configurations discussed above have been designed by experimental iterations. However, analyses of such configurations based on equivalent transmission line model [23] as well as based on full-wave spectral domain approach [24,25] have been reported recently. In [23], the transmission line model analysis of a two-layer rectangular patch is described. The upper patch is smaller, and the region common to both the patches is modeled as a coupled line section. Conceptually, the approach is similar to that described in [26] for a suspended patch antenna excited by an electromagnetically coupled inverted microstrip feed. For the configuration described in [23], a BW of 13.5% ($f_0 = 13$ GHz, VSWR < 2, lower and upper substrates have h = 0.8 mm and $\varepsilon_r = 2.55$) has been reported. For comparison the bandwidth for a single patch is only 4.46%.

Hankel transform domain analysis of two-layer circular patch configuration reported in [24,25] is an extension of the analysis of a single layer open circular microstrip radiating structure discussed in [27]. Complex resonant frequencies of the unloaded two-patch configuration are found by solving the equivalent transverse transmission line circuits for TE and TM components in the Hankel transform plane. Basis functions chosen for current distribution take into account the edge condition. Green's functions in spectral (Hankel transform) domain are in algebraic form. Coefficients of the immittance matrix are obtained via complex integration in the spectral domain. Equating coefficients determinant to zero yields complex resonance frequencies. Input VSWR characteristic is obtained by using natural complex frequencies computed and an equivalent circuit model. Radiation patterns are computed without inverting the Hankel transform current distribution. Numerical resonant frequency and VSWR variation are shown to be in good agreement with the experimental data ($\varepsilon_r = 2.55$, f ~ 2.4-2.8 GHz, dielectric layers 1.6 mm thick, air spacing between layers 10.0 mm, lower patch radius 20.8 mm, upper patch radius 21.0 mm).

Vertically stacked multi-resonator configurations are the most widely used broadband microstrip antenna elements. But, convenient practical design and analysis procedures are not available so far. Analytical treatments based on the extension of cavity and multiport network models are likely to be developed in the near future.

5. IMPEDANCE MATCHING NETWORKS FOR BROAD-BAND MICROSTRIP ANTENNAS

Approaches for widebanding of microstrip antennas discussed so far are based on either the concept of decreasing the resonator Q-factor, or on the concept of coupled multiresonator circuits. An entirely different approach makes use of wideband impedance matching networks to reduce the reflection loss at the input of the resonant patch antenna. It is an obviously natural solution from microwave circuit designers' point of view. However, a very small number [28-32] of microstrip antenna designers have looked upon the broadbanding problem from this angle.

As pointed out in [28,29,32], the maximum bandwidth obtainable by using an impedance matching network can be calculated by using Fano's broadband matching theory [33]. Treating the microstrip patch antenna as a parallel resonant circuit (as discussed in Sec. 2), it can be shown the input impedance bandwidth BW is given by

$$BW = \frac{1}{Q} \sqrt{\left\{\left(\frac{s}{GZ_0} - 1\right)(s-1)/s\right\}} \tag{6}$$

where G is the shunt conductance in the parallel resonant network and Z_0 is the characteristic impedance of the input transmission line. When $Z_0 = 1/G$, the relation (6) reduces to (2) of Section 1. However, one can find a value of Z_0 that will maximize the BW in (6) and this is given by

$$Z_0 |_{optimum} = \frac{2}{G(s + 1/s)} \tag{7}$$

For this optimum value of Z_0, the optimum BW is given by

$$BW_{opt.} = \frac{1}{Q} \sqrt{\left\{\frac{s^2-1)(s-1)}{2s}\right\}} \tag{8}$$

For the usual value of s=2, $BW_{opt.}$ given by (8) is about 1.2 times the bandwidth given by (2). It implies that if we do not insist on a perfect match at the center frequency, the frequency range over which the input VSWR is less than 2 can be increased by about 20%. This is consistent with Fano's theory [33] according to which maximum BW over which a parallel resonant circuit can be matched for a VSWR < s, is given by [28],

$$BW_{max} = \frac{1}{Q} \frac{\pi}{\ell n\{(s+1)/(s-1)\}} \tag{9}$$

Comparing the values given by (9) with those given by (2), it can be shown that a bandwidth improvement by a factor of 4.04 can be obtained (for s = 2) by designing a perfect impedance matching network.

Experimental results reported in [28] point out a bandwidth improvement by a factor of 3.9 (3 GHz, BW = 10%, s = 1.9) by using a shielded microstrip matching network consisting of two resonator elements etched on the same substrate as the patch. Configuration reported in [29], on the other hand, makes use of the feed probe inductance (occurring in a thick substrate coaxial-fed microstrip antenna element) as a part of impedance matching network. A series capacitance located at the top end of the probe (Kapton layer shown in Fig. 9) forms a series resonant circuit which is used for broadband matching of the parallel resonant circuit constituted by the resonant patch antenna. Experimental results show a marked improvement in the return loss (~11.3 GHz, ε_{r1} = 1.06, h_1 = 1.6 mm, ε_{r2} = 3.3 mm, h_2 = 0.15 mm) but bandwidth improvement data is not reported in [29].

Another wideband circular disc antenna making use of a series resonant circuit (feed probe inductance being a part of the series resonance) is described in [30]. The matching network (details not given) was optimized using CAD techniques and interfaced directly with a triplate array feed network. For a 6 mm radius patch (ε_r = 2.2, h = 3.2 mm) a bandwidth of 35% of VSWR < 1.5 has been reported. Radiation patterns over the whole band (7 to 10.5 GHz) shows little variation. Gain is about 6 dBi.

A forthcoming paper by Pues and Van de Capelle [32] describes in detail the transmission-line matching network design for broadband microstrip antennas.

A single stub impedance matching technique for increasing the bandwidth of an electromagnetically coupled microstrip patch (Fig. 10) is discussed in [31]. In this case, the input impedance of the electromagnetically coupled patch was measured by a network analyzer and component values for a network model (a capacitance in series with a parallel RLC) were obtained by a computer-aided model fitting procedure. A stub circuit was then designed to match to this load. A bandwidth of 13% (VSWR < 2, 3.375 GHz to 3.855 GHz, $\varepsilon_{r1} = \varepsilon_{r2} = 2.2$, and $d_1 = d_2 = 0.158$ cm) was thus obtained.

As evident from the above discussions, broadband impedance matching networks provide a really effective means of increasing the usable bandwidth of microstrip antennas. A combination of this technique with the other approaches discussed earlier would perhaps lead to the optimum configuration for wide bandwidth microstrip patches.

Electromagnetically Coupled Microstrip Patches

The example of broadband patch discussed above (described in [31]) makes use of an electromagnetic coupled patch (Fig. 10) consisting of two dielectric layers. The feeding microstrip line is printed on the lower layer and the patch is on the top of the upper dielectric layer. Such a coupling arrangement is frequently used in printed microstrip antennas [18, 26, 35-37]. Electromagnetically coupled patches are preferred as the feeding microstrip line network and other circuitry are located on a thinner (and possibly higher ε_r if desired) substrate and the radiating resonant patch is fabricated on the thicker and low ε_r layer.

Electromagnetically coupled patches are very convenient from the broadbanding point of view also. Impedance matching networks may be located on the lower layer underneath the patch without requiring additional substrate area. Also the capacitive coupling between the patch and the microstrip line constitutes a series capacitance which can be utilized as a component of the impedance matching network.

Some of these considerations are implicit in the bandwidth enhancement method for electromagnetically coupled strip dipoles discussed in [36]. In this paper, it is demonstrated that if parasitic metallic strips are incorporated in the structure either co-planar and parallel to the embedded microstrip line open-end, or between the transmission line and microstrip dipole, then substantial bandwidth enhancement can be obtained.

6. BANDWIDTH CONSIDERATIONS FOR CIRCULARLY POLARIZED PATCHES

Two different design philosophies are used for circularly polarized (CP) microstrip patches. One of these makes use of the two spatially orthogonal modes of a square or a circular patch. These two modes (each of them giving rise to a linearly polarized radiation when excited independently) are fed through two orthogonally located feeds by two signals which are equal in magnitude but are in phase quadrature. The bandwidth of these two-feed circularly polarized antennas is the same as that for the corresponding linearly polarized antennas, and all the broadbanding techniques discussed above are applicable to this class of circularly polarized antennas also. Reference [38] is an example where two vertically stacked circular resonator configuration is used to obtain a wide-band circularly polarized patch. A branch line hybrid is used to split the incoming signal for two orthogonally located feeds. A bandwidth of 13% (VSWR < 1.5) is obtained when the two patches (fabricated on $\varepsilon_r = 2.55$, 1.6 mm thick dielectric layers) are separated by air gap of

$0.08 - 0.09\ \lambda_0$. The radius of the upper patch is 1.0 to 1.1 times that of lower patch (frequency ~ 2.4 to 2.8 GHz).

The second design approach used for circularly polarized patches does not require a signal splitting network and makes use of only a single feed patch. There are several configurations [39-42] based on this approach. All of these configurations make use of two spatially orthogonal modes of the patch. The resonance frequencies of these two modes are made slightly different from each other. The patch is fed at a frequency in between these two resonances and the feed location is selected such that the two modes are excited equally. Since the resonance frequency of one of the modes is lower than the signal frequency and that of the other mode is higher than the signal frequency, a proper selection of frequencies causes the two modes to be excited 90° out of phase. Thus the conditions for a CP radiation are met.

The bandwidth of the single feed CP patches is limited not by the input VSWR (as is the case for linearly polarized patches), but by the degradation of the axial ratio as one moves away from the center design frequency. However, an analysis based on the cavity model [43] shows that the required separation between the two resonance frequencies and hence the axial ratio bandwidth is inversely proportional to the Q-factor of the patch.

A comparison of the axial-ratio bandwidths for various CP patches, given in [40], shows that the square ring configuration has a wider bandwidth (5.2%, axial ratio < 6 dB, f = 3.0 GHz, $\varepsilon_r = 2.50$, and h = 0.159 cm) as compared to crossed strip, almost square patch, corner chopped square patch, and a square patch with a diagonal slot configurations. Another wideband single feed CP configuration is annular ring with an ear configuration discussed in [42]. In this case, TM_{12} mode is used and higher bandwidth is obtained for small ratios of the outer to inner radii of the ring. A 6% bandwidth (axial ratio < 6 dB, $\varepsilon_r = 2.52$, h = 0.159 cm, f ~ 4.5 GHz) is reported.

As discussed in Section 2, rectangular ring and circular ring patches provide the maximum impedance bandwidth among the single-layer, single-resonator patch configurations. Thus it is not surprising to note that these two configurations also provide the maximum axial ratio bandwidth when used as single-feed CP configurations.

7. BROAD-BAND MICROSTRIP ARRAY CONFIGURATIONS

As the individual microstrip patches exhibit antenna gains of the order of 4-6 dB only, they are frequently used in linear and two-dimensional array configurations. As a general rule, arrays of broadband microstrip patches yield wide bandwidth. This is illustrated in [38] where 4x4 array of broadband patches is shown to yield 8.5% bandwidth (VSWR < 1.5, gain > 18.8 dBi, efficiency > 62%, SLL < -22 dB) at 2.6 GHz using the broadband CP patches discussed earlier.

Specially designed broadband configurations for linear arrays have also been reported [44,45]. The configuration discussed in [44] is a series fed linear array of rectangular elements. The individual elements of this array are two-port rectangular patches. For two-port patches, the VSWR bandwidth depends not only on the patch but also on the power transmitted to the output port. Thus much wider impedance bandwidths are available in series-fed array. The design reported in [44] has a bandwidth of 40% (VSWR $\leq$ 2) and a gain of ~10 dB at x-band. The configuration employs a double section non-uniform antenna (Fig. 11) having a length equal to about 5.0 λ_{max} and consisting of two rows, spaced about 1.5 λ_{max}, each of seven radiating patches.

The application of the log-periodic technique to the series-fed electromagnetically coupled overlaid-patch array has been discussed in [45]. A k-ß analysis of microstrip arrays is reported and indicates that the microstrip patch is not an optimum element for log-periodic arrays and that elements connected directly to the feed line result in the arrays having a limited bandwidth. The addition of the series capacitance to the patch equivalent circuit, implemented by electromagnetic coupling, allows an optimum to be approached. Log-periodic overlaid patch array design (Fig. 12) has been reported to have a 4:1 bandwidth with an input return loss of 8 dB, a gain of 8 dBi and a 30° backfire beam whose beamwidth varies from 63 to 32 degrees across the band. In spite of limitations on the ultimate BW obtainable, this design certainly extends the useful application areas of microstrip antennas.

8. CONCLUDING REMARKS

Various techniques used for increasing the bandwidth of microstrip patch antennas have been reviewed. It is pointed out that the consideration of the *impedance bandwidth* alone is not sufficient for broadband microstrip antennas, and a concept of *radiation bandwidth* is introduced.

It has been discussed (in Sec. 2) that the Q-factors of the resonant patches can be decreased and hence their bandwidths can be increased by increasing the height of the dielectric substrates. The problems associated with thick substrates are: increased reactance of the feed junction, and the radiation from the microstrip circuits included on the same substrate. These difficulties may be overcome by using electromagnetically coupled patches with feed network and associated circuitry on the lower (and thinner) substrate. There is a need to develop design procedures for electromagnetically coupled patches based on the extension of cavity model or multiport network model.

Among the three other approaches used for obtaining wider bandwidth, more attention needs to be directed to the application of wideband impedance matching network approach discussed in Sec. 5. The only difficulty with this method is the need for larger substrate area required for incorporating the matching network. Again, the use of electromagnetically coupled patches should prove very useful since the matching network can be incorporated at the lower level, possibly underneath the patch itself. Further research efforts need to be pointed in this direction.

Between the two different methods of constructing multiresonator coupled patch configurations, the two-layer configuration with vertically-stacked patches requires smaller area and does not suffer much from pattern degradation with frequency. This configuration is used frequently but most of the designs are carried out by experimental iterations. Convenient models and analysis procedures need to be developed so that the design and optimization of these configurations can be carried out more systematically. Among the various configurations with multiple patches on the same layer, the configuration with the six narrow patches coupled along the non-radiating edges [15] has yielded the most promising results. Further research leading to analysis and design of this type of geometry is desirable. A multiple coupled microstrip line model, proposed recently for two patches coupled along the non-radiating edges [1], appears to be a reasonable approach for this purpose.

The search for an "ideal" wideband printed microstrip antenna is still on. Perhaps a combination of various approaches discussed in this review would lead to an optimum broadband configuration. We can look forward to continued research in this area.

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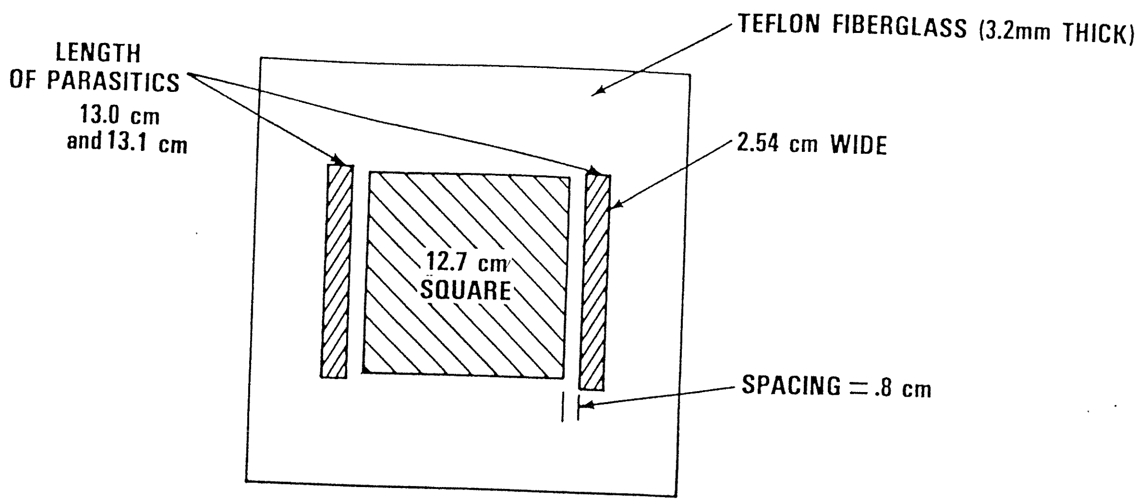


Fig. 1 A rectangular patch antenna with parasitically coupled strips [10].

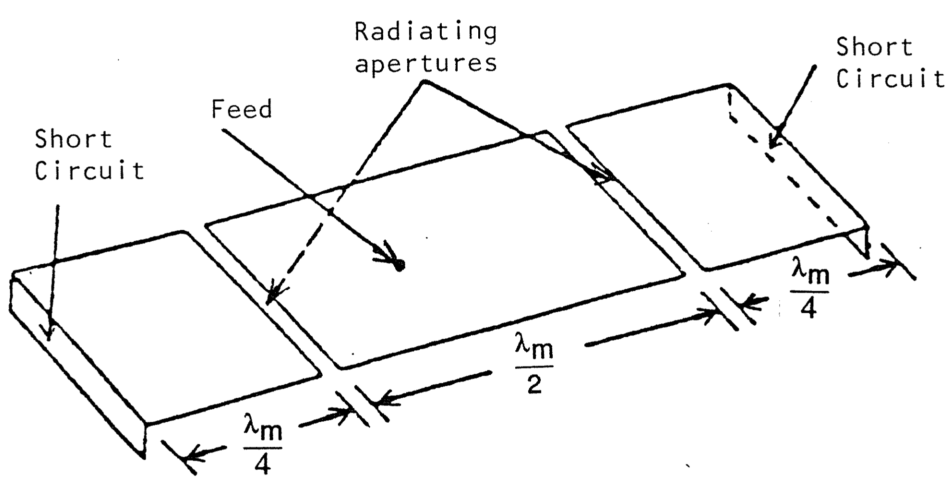


Fig. 2 A rectangle with two λ/4 shorted patches coupled to the non-radiating edges [11].

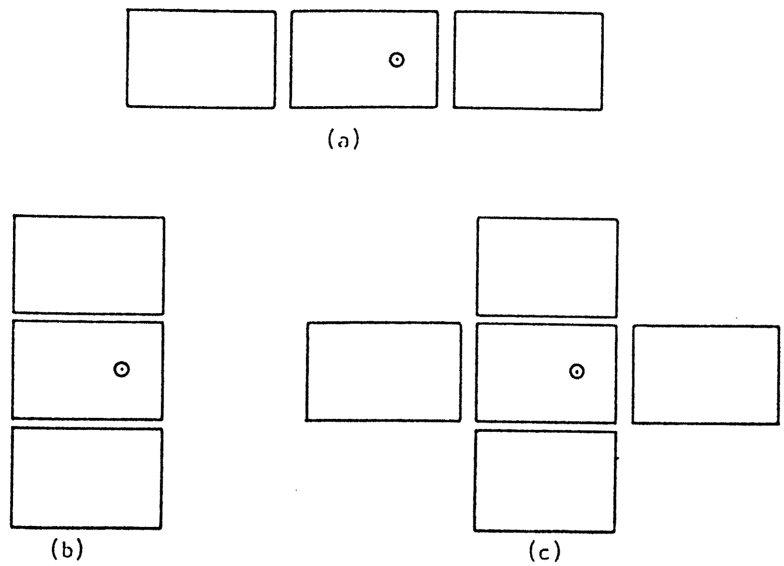


Fig. 3 (a) Radiating-edges gap-coupled microstrip antenna [12];

(b) Non-radiating edges gap-coupled microstrip antennas [13]; and

(c) Four-edges gap-coupled microstrip antenna [13].

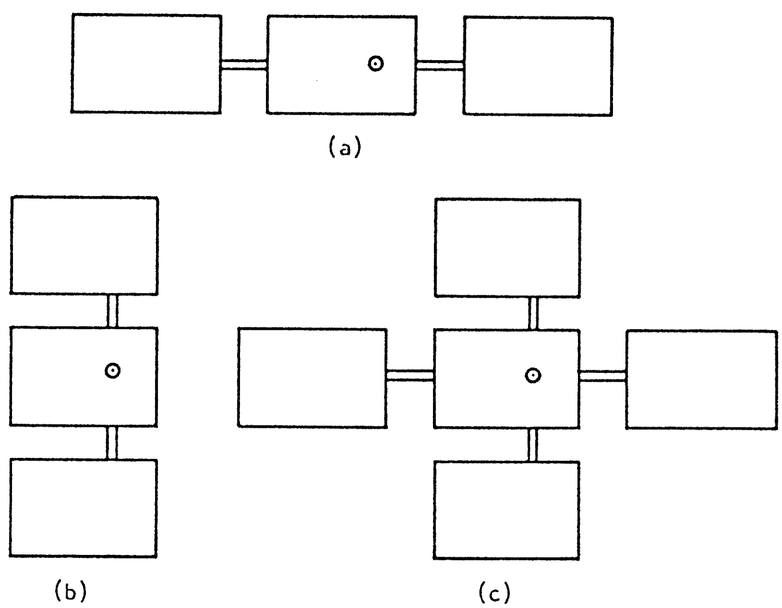


Fig. 4 (a) Radiating-edges directly-coupled microstrip antenna [14];

(b) Non-radiating edges directly-coupled microstrip antenna [14] and

(c) Four-edges directly-coupled microstrip antenna [14].

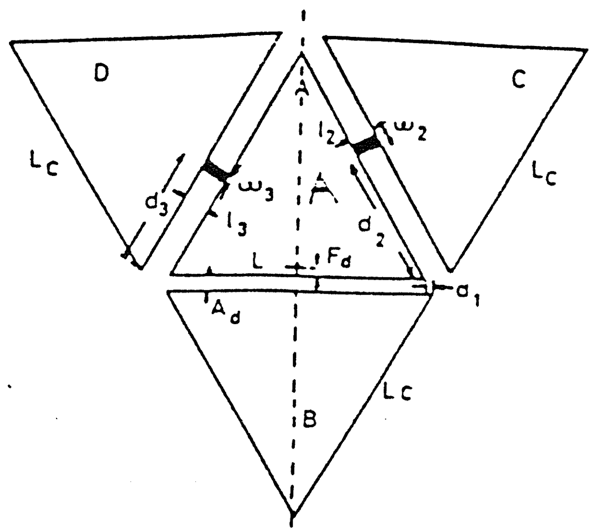


Fig. 5 A multiresonator microstrip antenna using four triangular patches [15].

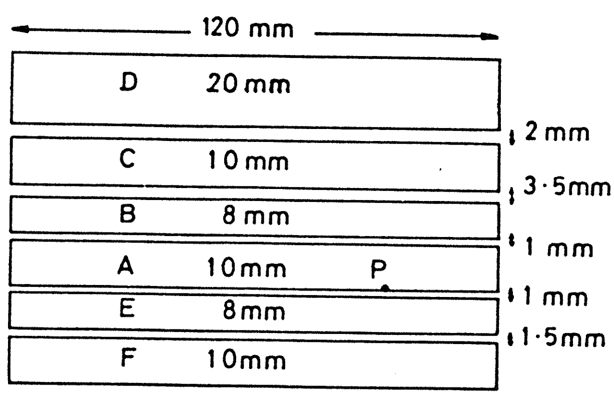
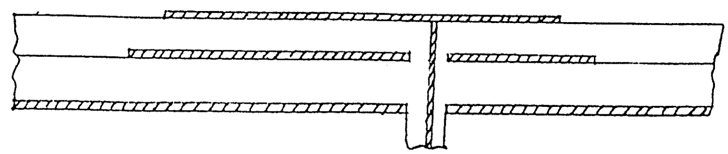
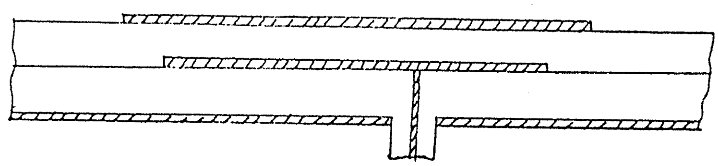


Fig. 6 A multiresonator microstrip antenna using six narrow patches coupled at the non-radiating edges [16].



(a)



(b)

Fig. 7 (a) Two-patch two-layer antenna with smaller patch located on the top layer.

(b) Two-patch two-layer antenna with larger patch located on the top layer.

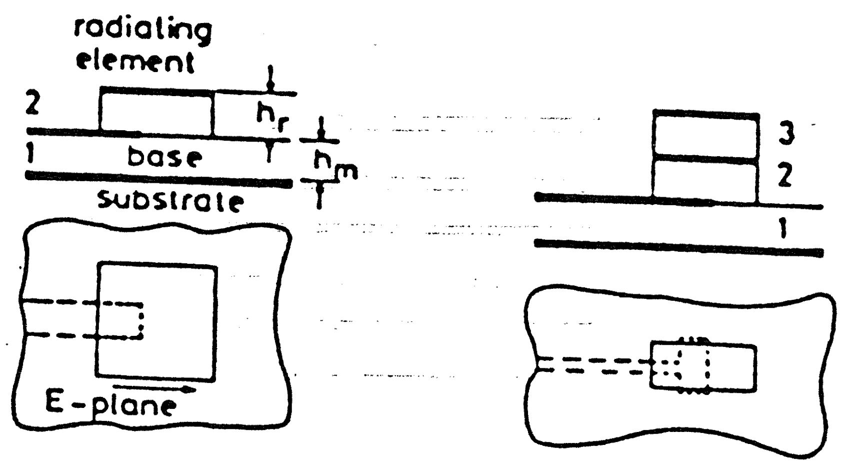


Fig. 8 (a) and (b) Examples of wideband microstrip antennas for circuit integration [18].

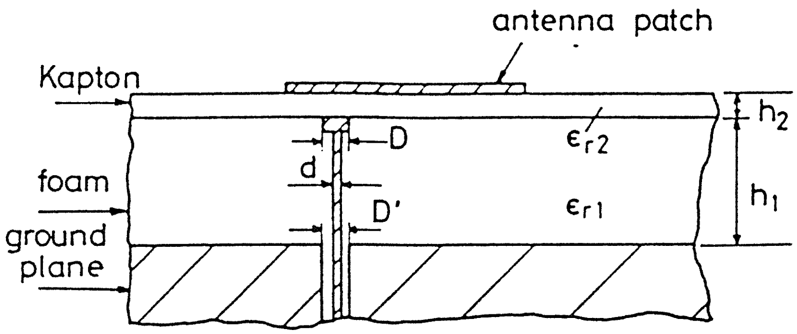


Fig. 9 A microstrip antenna with a series L-C network used for wideband impedance matching [29].

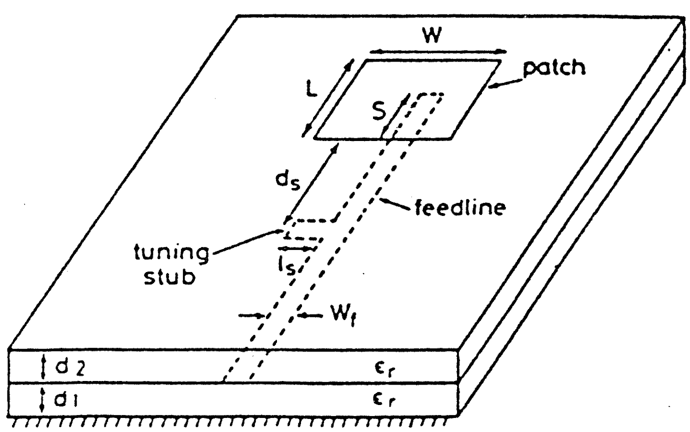


Fig. 10 A single stub impedance matching technique for increasing the bandwidth of an electromagnetically coupled microstrip patch [31].

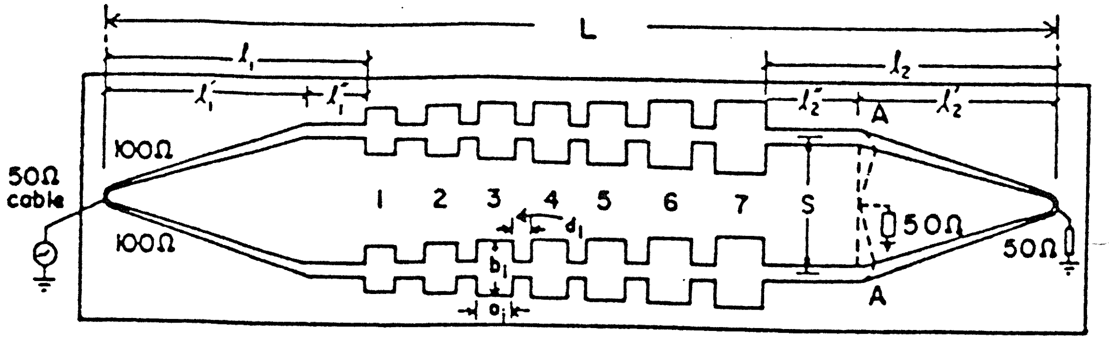


Fig. 11 Double-section wideband series-linear array of microstrip patches [44].

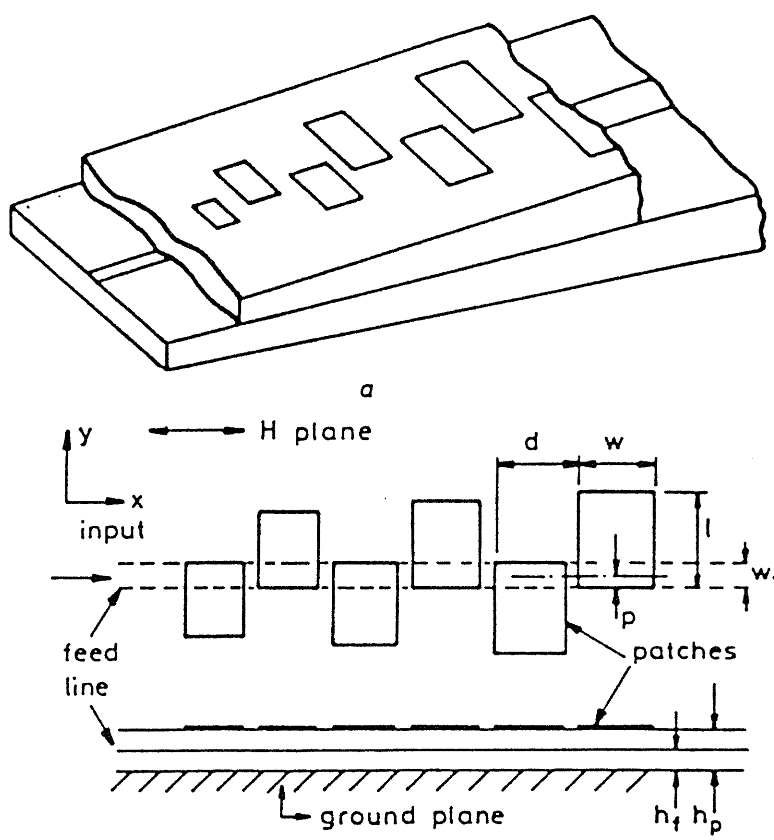


Fig. 12 A log-periodic overlaid array of electromagnetically coupled microstrip patches [45].