Investigation/design of scalable RF metaresonator for building material

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2019

Sum, Y. L. (2019). Investigation/design of scalable RF metaresonator for building material.
Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/90132>

<https://doi.org/10.32657/10220/48427>

**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**INVESTIGATION/DESIGN OF SCALABLE RF
METARESONATOR FOR BUILDING MATERIAL**

SUM YEE LOON

**SCHOOL OF ELECTRICAL AND ELECTRONIC ENGINEERING
2018**

INVESTIGATION/DESIGN OF SCALABLE RF METARESONATOR FOR BUILDING MATERIAL

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School of Electrical and Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfilment of the requirement for the degree of
Doctor of Philosophy

2018

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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Supervisor Declaration Statement

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

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[A/P Soong Boon Hee]

Authorship Attribution Statement

This thesis contains material from 2 papers published in the following peer-reviewed journal(s) where I was the first and/or corresponding author.

Chapter 3 and 4 are published as Y.L. Sum, V. Rheinheimer, B.H. Soong and P.J.M. Monteiro, "Scalable 2.45 GHz Electrically Small Antenna Design for Metaresonator Array," in *The Journal of Engineering*, vol. 2017, no. 5, pp. 170-174, 5 2017.

The contributions of the co-authors are as follows:

- Prof Monteiro and Prof Soong reviewed the manuscript drafts and provided research guidance in the interpretation of results.
- I prepared the manuscript drafts. The manuscript was revised by Dr Rheinheimer.
- I designed the study and performed all the laboratory work at the School of Electrical and Electronic Engineering. I also analyzed the data.
- All design and experiment were conducted by me in INFINITUS, NTU.
- Dr Rheinheimer assisted in the conduction of the experiment.

Chapter 5 is published as Y.L. Sum, V. Rheinheimer, B.H. Soong, P.J.M. Monteiro, "Effects of Cement Paste Enhanced with Iron-Based Magnetic Particles on an Embedded Small Resonator Antenna," *Nature Scientific Report*, 7, Article number: 15185, 2017.

The contributions of the co-authors are as follows:

- Prof Monteiro and Prof Soong edited the manuscript drafts and provided research guidance in the interpretation of results.

- I prepared the manuscript drafts. The manuscript was revised by Dr Rheinheimer.
- I co-designed the study with Dr Rheinheimer and performed all the laboratory work at the School of Electrical and Electronic Engineering and University of California, Berkeley. Dr Rheinheimer and I also analyzed the data.
- All design and experiment were conducted by Dr Rheinheimer in INFINITUS, NTU and University of California, Berkeley.
- Dr Rheinheimer assisted in the conduction of the experiment.

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Acknowledgements

First, I would like to express my greatest gratitude to my supervisor, Prof. Soong Boon Hee of Nanyang Technological University, NTU, for his advice and inspiration given throughout this doctoral journey. I am also grateful for the opportunity given to be part of the Singapore-Berkeley Building Efficiency and Sustainability in the Tropics, SinBerBEST Program. This program provided me an international stage to work with experts from different fields in the green building context.

I am equally grateful to Prof. Tseng King Jet of Singapore Institute of Technology, SIT for his continuous encouragement and trust in supporting my ideas throughout this time in and outside of the SinBerBEST program. I would also like to thank Dr. Liu Wei of Singapore Institute of Manufacturing Technology, SIMTech A*STAR for his kind help whenever I needed them. I am also very fortunate to be able to work with Prof. Paulo J.M. Monterio of University of California, Berkeley, UC Berkeley whose insightful analysis helped me tremendously in my research. The research at Berkeley also helped enrich my academic experience.

My fellow co-workers in SinBerBEST and NTU also provided exceptional help throughout my research. Especially thankful to Dr. Edwin Chan Yiu Wing, who provided ideas and guidance in my work. He is also instrumental in the success of the many projects that we embarked on over the years. Special thanks go to Dr. Vanessa Rheinheimer who spent hours on end discussing and working in the laboratories producing samples for our research and

experiments. Mr. Benjamin Chew also helped me throughout this journey in thinking through many of the challenges encountered. Mr. Ng Teng Kwee of INFINITUS, NTU has also provided essential technical support in the laboratories.

I would also like to gratefully acknowledge SinBerBest in funding much of my Ph.D. work.

I am especially thankful to my mother, Diana and sister, Hui Ting for their patience and encouragement in the pursuit of this research. They are always ready to lend a hand whenever I needed help. I would also like to specially thank my wife, Jasmine for sharing the joys and pains through the ups and downs that we experienced together. Finally, and most importantly, I am grateful to have an understanding daughter, Fiona, who waited patiently over these years. Fiona is the best daughter any father can ask for.

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Abstract

Globally, governments are building smarter, greener and more sustainable buildings with the help of technological innovations. Deploying embedded WSN in buildings is one of the promising technologies that enables such a vision. However, such sensors face challenges in sourcing long term power, and transmission of wireless signals in and out of building materials. One way to source power is from the ambient RF signals. To harvest such energy, RF energy harvesting antenna is an important component in embedded WSN systems. This antenna can also be responsible for wireless communications. In addition, building materials enclosing the embedded antenna also affects the transmission of RF power for harvesting and communications.

In this thesis, a study on the concepts of metamaterial and electrically small antennas is performed to combine both concepts in the design of an electrically small antenna, and a scalable antenna array suitable for RF energy harvesting and wireless communications. Another aspect affecting RF energy harvesting and wireless communications indoors is also investigated in this thesis. Analysis into some commonly used admixtures to compare the effects on the embedded arrays and the resulting materials is done.

Combining concepts from metamaterials and electrically small antenna, an electrically small metaresonator is designed. This antenna has a maximum dimension of one-tenth the wavelength of WiFi spectrum at channel 13. Using the stubbed design, the resonant frequency can be tuned to the required centre frequency at channel 7. The resulting antenna

has a S_{11} of -25 dB, gain of -16 dB and an efficiency of -9 dB, within expectations of an ESA.

This electrically small metaresonator is then used as an array element to form into an array using DC combiner topology. Due to its geometry and combining the concepts of metamaterials, the resulting metaresonator array achieves close cell separation distance of 0.0165λ . The S_{11} of this array is also able to achieve -25 dB, a gain of -4 dB and an efficiency of -6 dB. Furthermore, an analysis of feeding the array at different feed points also demonstrates the flexibility of this array, making it suitable for embedded WSN in buildings.

Next, an investigation into the common building material, cement is done. Comparing admixture pigments typically used in the building industry for colouring of mortar and concrete, the results indicate that cement paste enhanced with 4% micro-sized particles of iron (III) oxide 1) improves the S_{11} of the embedded array by as much as 20 dB, 2) increases its bandwidth while maintaining the overall shape of the S_{11} , and 3) produces a lesser shift in resonant frequency than cement paste alone. Further analysis into S_{21} shows that adding micro-sized iron (III) oxide into cement paste improves the RF power transmission into the material compared to cement paste only sample by as much as 10 dB.

Further investigations into the application of plaster is done to develop an understanding of certain applications in buildings. The results obtained shows that plaster samples are the same or worse than cement paste samples. Furthermore, the findings suggest that embedding the metaresonator array in between a layer of plaster (enhanced or not) and mortar is not suitable for wireless applications as the S_{11} is significantly distorted to an unacceptable range. Analysing S_{21} , it is also observed that plaster enhanced with iron (III) oxide layered on mortar also does not significantly change the RF transmission property of the material. These findings indicate that embedding the metaresonator array in plaster or plaster layer on mortar is not suitable for embedded WSN applications.

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List of Abbreviations

BCA	Building and Construction Authority
BIM	Building Information Modelling
CMA-ES	Covariance Matrix Adaptation Evolution Strategy
CSRR	Complementary Split Ring Resonators
CST	Computer Simulation Technology
EH	Energy Harvesting
EIRP	Equivalent Isotropically Radiated Power
EM	Electromagnetic
ESA	Electrically Small Antenna
FDTD	Finite-difference time-domain
IEEE	Institute of Electrical and Electronics Engineers
IMDA	Info-communications Media Development Authority
IOT	Internet of Things

IPC	Association Connecting Electronics Industries
ISM	Industrial, Scientific and Medical
LEED	Leadership in Energy and Environmental Design
MM	Metamaterial
MOM	Method of moments
PCB	Printed Circuit Board
RF	Radio Frequency
RFID	Radio-frequency identification
SE	Shielding effectiveness
SHM	Structural Health Monitoring
SRD	Short-range devices
SRR	Split Ring Resonators
USGBC	United States Green Building Council
WSN	Wireless Sensor Network

Chapter 1

Introduction

1.1 Background

The world's population is increasing rapidly since the industrial revolution in the 1800s. Currently, the percentage of people living in the urban environment is also larger than those living in the rural areas. By 2050, the projected urban population will increase by another 2.5 billion, with 66% of us living in the urban areas [1]. This trend of human activities places increasing pressure on building designers, engineers, and operators to find sustainable ways to accommodate occupants in buildings, not only for the occupants, but also for the environment.

In recent years, governments worldwide have established standards for the building industry to create green buildings for both occupants and the environment in a healthy, efficient, and cost-effective way. For example, in the United States, the United States Green Building Council, USGBC in less than twenty years, has established a set of credit system certification. This system awards credits to buildings that met certain criteria ranging from transport, building materials, water monitoring, air quality, and indoor light pollution [2]. The accumulated credits determine the category of Leadership in Energy and Environmental Design, LEED certification starting from certified, to silver, gold, and platinum. Another example, in Singapore, the Building and Construction Authority, BCA also has a set of standards, named Green Mark for

the building industry to follow. Similar to LEED, this is a guideline on how credits are awarded to the different categories including thermal comfort, spatial quality, and smart operations [3]. In other countries, such as Australia, Emirates, Croatia, and Sri Lanka, they too have their own green building councils to address the growing green and sustainable building needs.

To meet the various criteria set out in these certifications, using sensors is increasingly important, and has been properly recognised in these standards. For instance, Green Mark awards credits for the use of sensors such as temperature, carbon dioxide and illumination sensors. In the future, the use of sensors together with internet of things, IOT will be even more widespread as the cost and energy consumption of the sensors come down. According to an estimate done by Deloitte, the compounded growth rate is 78.8% from 2015 to 2020, totalling to a combined 1.3 billion sensor deployments [4]. Recognising this trend, many governments are proposing and implementing many programs and initiatives in their cities to promote the concept of smart cities. Some of the remarkable cities and projects are Smart Nation initiative by Singapore [5], Amsterdam Smart City in Netherlands [6], iBarcelona in Spain [7], MK:Smart in Milton Keynes, Buckinghamshire, London [8] and Smart Cities Mission in India [9].

In the anticipation of powering these sensors and wireless transmission in the concept of IOT in buildings, numerous research have been carried out ranging from energy efficient routing of wireless data [10][11], to low power sensors and electronics in the micro-watt range [12][13].

1.2 Energy Harvesting for Indoor Sensors

It is a certainty that more sensors will be deployed in existing and future green buildings. In these indoor spaces, some of the common sensors installed are occupancy sensors [14], fibre optic sensors [15], with some embedded into building structures such as moisture sensors [16], stress-strain sensors [17], and even temperature sensors [18] in nuclear facilities. To power these sensors, traditionally, either power grids are laid to provide AC or DC power, or local energy storage like batteries are supplied to the sensors. In recent years, and more so in the future, power sources will come from the environment. For indoor spaces, the possible energy

sources can be from ambient light [19], mechanical vibrations [20][21], thermal energy [22], electric field from medium and high voltage power lines [23], and radio waves [24].

Typically among the many sources of energy, conversion from kinetic energy such as push buttons, heel strike and vibrations, and light energy [25][26], has the highest amount of power and energy that can be harvested in the range of Watts to milli-Watts. Although mechanical sources of energy such as push buttons provide large amount of power, the amount of power that can be harvested are reduced as they are increasingly being replaced in modern day buildings with other types of switches like buttons with lower travel distance and touch enabled buttons. Another factor to consider during implementation and deployment of wireless sensors is the position of the sensors. When designers and engineers try to deploy sensors as unobtrusive to the occupants as possible, they are usually installed in places hidden from sight, or placed at a certain height above the reach of occupants, sometimes near the ceiling or above false ceiling. At places like these, challenges exist to integrate with the mechanical devices that are more assessible at human heights. In situations where these sensors are provided with a high-power source, they cannot readily consume this energy. If the source of energy is instantaneous such as push button or heel strike, then there is a problem of designing an energy storage that can absorb this energy in a very short time. In another situation when energy is not present for an extended duration of the day such as solar or indoor artificial lighting, a large energy storage that can last throughout this duration is required. In either scenario, significant effort in design of the energy storage system is required.

Compared with the common kinetic or light energy sources, ambient RF energy have the least amount of power at less than $1\text{ }\mu\text{W}$ [27]. For low power sensors, which do not require high power to operate, this amount is sufficient for their function of sensing with careful design. In the case of ambient RF sources, although the power available is small compared to the other sources, this energy can almost be available all the time if the spectrum is selected correctly. At places not easily accessed by humans, RF energy harvesting, RF EH is a good method of sourcing power. As wireless signals can propagate throughout the entire indoor space, and even near ceiling heights, wireless sensors scattered throughout this space is within reach of this

ambient source of energy. If the power available is not sufficient, a capacitor can be used to store up enough energy before activating the sensors. Compared to other sources, the sensor nodes using RF EH do not require large batteries or be restricted to the places of deployment.

Other than power levels, energy conversion efficiency is also another consideration. For the many different energy harvesting methods, the typical energy conversion efficiency is between 10% to 20% for solar cells, to about 50% for heel strike [27]. Comparing these to RF-DC energy, the conversion efficiencies ranges from 40% [167] to more than 95% [168]. This higher efficiency makes ambient RF a possible power source for RF EH based sensors.

In the context of buildings, RF EH in buildings is getting popular due to the growing number of devices that rely on wireless communications for data transmissions and controls. This is partly due to recent developments in wireless sensor and control nodes in buildings [23]—[31]. Together with more people who bring your own devices (BYOD) to the office [32]—[34], the RF spectrum is getting more crowded. For RF EH applications, this crowded spectrum is beneficial to the amount of energy that can be harvested. Among the commonly used spectrum from the kilo-hertz range used by audible radio broadcast, to mega-hertz and sub giga-hertz range used by television broadcast and 2G mobile networks, to less than 3 GHz range used by modern day mobile networks and 2.4 GHz Industrial, Scientific, and Medical, ISM band, the most common spectrum used in buildings is the unlicensed 2.4 GHz ISM band. This band has a good balance between coverage, bandwidth, cost, and the size of electronics and antennas. Due to these considerations, many protocols used in buildings such as WiFi, ZigBee, Bluetooth, and a range of non-standard proprietary protocols use this band. As more sensors are wirelessly connected, this spectrum will be get even more crowded, thus making more ambient RF energy available for harvesting.

When deployed, RF EH wireless sensors nodes, WSN, can achieve “deploy and forget” implementation. If the right self-organising wireless protocol such as Zigbee or Bluetooth mesh is used together, the nodes can be scattered in many places, and system integrators will not need to worry about either powering the nodes or conducting time consuming network planning

to link all the nodes. Therefore, as more devices are connected wirelessly in buildings, RF EH will potentially be a more practical way to drive low-power WSN.

1.3 Embedded Sensors for Building Applications

We now know that to make buildings smarter, greener, and sustainable for both humans and the environment, it is undeniable that self-organising, and self-powering WSNs play an important role. Integrating these sensors well with the building design is the next challenge facing designers and engineers. Occupants perceive that buildings are smart when sensors exist without them knowing but elements within the building react to their actions. In addition, hidden sensors also allow better integration into the building right from the start of the buildings design phase. This is especially so with the use of building information model, BIM which contains information of the building from the initial design phase, to the operations, and final decommissioning of the building. These sensors can provide critical data related to the building such as structural health and changes to the building materials as it ages.

One way to hide sensors from the occupants is to embed them into surfaces such as walls, façade, ceilings, floors and partitions. By doing so, these sensors can be cast into precast concrete structural elements at the factory [35] before assembling on site. Concrete is primarily used in modern day structures due to its excellent resistance to water, plastic consistency to form into different shapes before solidification, and its low cost [36]. Modern day concrete is generally made up of four main components. First, hydraulic cement, or commonly used Portland cement is basically made up of reactive calcium silicates that turn into calcium silicate hydrates during the hydration process. The mixture of hydraulic cement and the second component, water forms hydraulic cement concrete. The third component is aggregate. There are four common types of aggregate. First, coarse aggregate are particles that are larger than 4.75 mm, second, fine aggregates are particles smaller than 4.75 mm but larger than 75 μm , next, crushed blast-furnace slag, and last, construction or demolition waste. When water, cement and sand (a type of fine aggregate) is mixed, mortar is formed. The final component typically used in modern day concrete is admixtures, which are materials that do not fall under

any of the three mentioned components. They are added immediately before and during the mixing.

Admixtures are usually added to modify or change the behaviour of the cement mix during the hydration process or the properties of final concrete product. The five classes of admixtures include air-entraining, water-reducing, retarding, accelerating, and plasticizer. Other admixtures whose functions include corrosion inhibition, shrinkage reduction, alkali-silica reactivity reduction, workability enhancement, bonding, damp proofing, and colouring are classified under special category [37]. For colouring of concrete, the common material used are cobalt oxide for blue, brown oxide of iron for brown, chromium oxide for green, and red oxide of iron for red [38]. Hence, embedded sensors used in buildings are generally encased in these materials.

Embedded sensors can provide useful information regarding behaviour of the building materials. In the pre-fabrication industry, a way of embedding these sensors into the concrete is as shown in Fig. 1-1. The sensor can be attached to the reinforcing bars, or rebar in the factory before the cement is poured. For example, Concrete Sensors implement temperature and humidity sensors solution in this manner during the curing stage of the concrete [39]. As the environment affects the curing time, these embedded sensors provide important information during the curing process such as revealing when the concrete is set. When the concrete is set earlier, additional time spent waiting can be saved for another batch of pouring. Furthermore, the embedded sensors can continue to operate for another few more years, providing more information that can be fed into BIM. However, as the batteries do not have a means to be recharged, the sensors cannot operate as long as the lifespan of the building.

An important note from this implementation is the distance that is advised by the manufacturer with respect to the depth of embedding the sensors. Due to the good RF absorbing nature of concrete, although the depth of the sensors can go up to a distance of 180 cm, the depth the transmitter can only be embedded up to one-ninth of that [39].

Fig. 1-1 Embedding sensors into building materials

Other types of sensors that can be embedded are sensors that can tell the structural health of the concrete. Such structural health monitoring, SHM sensors can be made up of ferromagnetic microwire [40], piezoelectric materials [41][42] and fibre-optics [43], which do not require a lot of power to operate. For longer duration of operations, it is therefore possible to power these embedded sensors with energy that can be harvested from the environment.

1.4 Objectives and Contributions

It is with high certainty that future green and sustainable buildings will use more wireless sensors to enable buildings to be smarter for both the occupants and environment. Enabling embedded sensors to transfer data and command wirelessly, while having a means to receive injected RF energy to sustain its operations or prolong the battery life are important factors for WSN in buildings. Although the power received from the external RF sources will be weak, this ability enables embedded sensors to have a long service life. The objectives of this thesis are, therefore 1) to design an antenna that can satisfy the demands of embedded WSN in a built

environment, and 2) investigate how certain admixtures used in building materials can be used to improve the properties of such materials and the antenna.

Antenna is an essential part of WSN in buildings for wireless communications and receiving injected RF power. For wireless communications in buildings, the type and design of this antenna affects the way embedded sensors can be deployed and implemented. Planar antenna is a type of antenna that is conformal enough to suit large flat surfaces such as walls, ceilings, floors, façade or partitions such that they can be integrated well with the interior design. In addition, as these surfaces can be of different sizes, it is also advantageous for this antenna to be able to change its size to suit different areas. If by scaling the area of the antenna such that certain qualities can be improved, then this will be an additional benefit for such applications. The next important feature of this antenna is the cost. Cost directly affects the adoption rate and the size of WSN. Lastly, when embedded or adhered to surfaces, this antenna does not have the flexibility of changing its orientation after installation. Thus, an antenna that has the ability to tune in some way after deployment is another important property to have.

Next, a suitable antenna for indoor embedded WSN is also one that can be used to receive injected RF energy. This dual function is beneficial for either prolonging the life of the battery, or in the case of sufficient energy received, totally replace the battery. A RF energy receiving antenna for indoors embedded WSN also share the four properties mentioned in the previous paragraph. The property of scalability is especially important for RF energy receiving antenna as the area of antenna directly affects the amount of RF energy it can receive. When applied to the context of indoor spaces, where flat surfaces are wide and numerous, an antenna that can scale up in area is especially useful to increase the amount of RF energy received.

Therefore, the first objective is to design a suitable antenna for wireless communications and RF energy receiving in embedded WSN that is planar, low cost, scalable, and tuneable.

With the knowledge that embedded WSN for the indoor environment will be implemented in the presence of building materials, it is important to investigate how different materials affect the antenna when it is embedded. It is commonly known that concrete structures in indoor

spaces negatively impact propagation of RF signals, especially those frequencies in the gigahertz range. Concrete structures such as walls, ceilings, floors and façade have high shielding effectiveness, SE, in addition to reflection and refraction causing multipath. For RF EH, concrete structures thus reduce the amount of energy the antenna can harvest. This is especially so when the antenna is embedded. However, when admixtures are added to the cement mix, which is frequently practiced in the building industry, little studies have been done to investigate the effects these materials have on EM propagation or how it affects embedded antennas.

It is therefore valuable to understand, which admixtures when added to the cement paste, is able to improve its EM properties such that it benefits the embedded antenna. The advantage of this is that, if such admixture exists, does not require extra certification for its suitability to use in buildings in terms of occupant and building safety. Also important is how the application of such material can affect both the material and embedded antenna.

Hence, the second objective of this thesis is to investigate some of the common admixtures of iron-based oxides used in building materials to establish an understanding how they affect the RF transmission properties of the materials, and the performance on the antenna. Using the results from this investigation, the concentration of this admixture is varied to study how the percentage affect the material and antenna. Finally, an additional study was done to study how the different common applications of such materials can alter their effects.

Aligning with the objectives identified, the research outcomes achieved in this thesis include a planar scalable antenna array that is low cost, and tuneable, and an analysis into the type and concentration of admixture that can improve the properties of enhanced concrete and the embedded antenna.

The major contributions are listed as follows:

1. Combining the concepts of metamaterial and antenna array, the maximum element size of one-tenth the wavelength is determined. When the element of this size is chosen, an antenna array can be designed with the elements placed closer together.
2. Using the concepts of metaresonators, a gigahertz planar electrically small metaresonator antenna is designed and fabricated using low cost materials and standard printed circuit board, PCB fabrication processes without passive component loading. The stubbed design allows fine tuning of resonant frequency and improves matching of the ESA.
3. Using results achieved in the above contributions, the electrically small metaresonator is used as a unit element to form a scalable metaresonator array. With the stubbed design, the resonating frequency was fine-tuned to the desired frequency by tuning the length of the stubs. In addition, due to the design of the feed points, the directivity of this array can be tuned by feeding at different feed points. It is also able to receive power from a dedicated RF source operating at the resonating frequency.
4. Being applied to indoor spaces, admixtures commonly used in cement mixes are investigated. Among the admixtures and their concentration tested, 4% concentration by weight of micro-sized iron (III) oxide is identified as having the best effects on the material and the embedded metaresonator array. It is also further analysed that applying this admixture in plaster on mortar does not produce positive effects compared to admixture in cement paste.

1.5 Organisation of Thesis

The rest of the thesis is organised as follows. In Chapter 2, a study is first done to select a suitable frequency spectrum for indoor RF EH. An overview of the previous work on RF EH antennas and arrays is presented. This is followed by a determination of a suitable element size for combining concepts of metamaterial and antenna array. Recognising one-tenth the wavelength is a suitable maximum dimension, a review on the electrically small antennas is

performed. Recent developments in metamaterial-based rectenna arrays for RF EH is then shown. Finally, building materials and their effects on EM propagation are covered.

Chapter 3 presents a proposed gigahertz ESA using the concepts of metaresonators. Compared with other structures, the 0.1λ proposed electrically small metaresonator is planar, fabricated on low cost FR-4 substrate and using standard PCB fabrication processes. Having the stubbed design, the ESA was able to fine-tune to the desired resonating frequency. Furthermore, this ESA can function as an element in an antenna array.

Chapter 4 demonstrates the possibilities of using the proposed 0.1λ ESA as an array element in an antenna array. Based on the concept of metamaterial, the distance of element separation is significantly lesser compared to conventional arrays. Using the stubbed design, the array was able to fine-tune to the desired resonating frequency. An analysis was then done to show that by changing the position of the feed point, this array can be tuned to change its directivity.

Chapter 5 investigates the effects of iron-based magnetic particles enhanced cement paste on the proposed metaresonator array embedded in it. These particles are typically used as admixtures to colour concrete during mixing. Not only are different types of particles investigated, the size of the particles is also tested. The findings show that micro-sized iron (III) oxide particles have the best effects on the EM transmission property of the material and the S_{11} of the embedded array.

Chapter 6 then studies the different concentration of micro-sized iron (III) oxide in enhanced cement paste to determine a suitable percentage by weight that can be used to affect the material and embedded array. Working within a range of admixture pigment weight percentage typically used in the industry, it is validated that 4% concentration of micro-sized iron (III) oxide has the most significant influence on the EM transmission property of the material and S_{11} of the embedded array.

Chapter 7 further compares two different but common applications of iron (III) oxide admixtures in the building environment. The comparison shows that the common application

of iron (III) oxide enhanced plaster on mortar does not have a positive impact on the EM transmission property of the material and S_{11} of the embedded array.

Finally, Chapter 8 concludes this thesis and provides recommendations for future work.

Chapter 2

Literature Review

2.1 Identifying Suitable RF Spectrum for Energy Harvesting Applications

To establish the optimal RF spectrum for ambient RF EH, first, a survey of the available spectrum is required. There are certain RF spectrums that are commonly used for communications. For example, in Singapore, Info-communications Media Development Authority, IMDA specifies the spectrum that can be used locally. An example is shown in [44] where the spectrum allocation for 3 kHz to 300 GHz is specified. Also, from the Radio Spectrum Masterplan [45] by IMDA, the spectrum allocations for wireless services are broadly divided into five categories, namely; mobile services, fixed services, satellite services, broadcasting services, and short-range devices, SRD. For the frequency ranges of each service, three levels of review priority are assigned. “High priority” spectrum bands have an indicative timeframe for spectrum review of 1 to 2 years, “medium priority” spectrum bands have 2 to 4 years and more than 4 years for “low priority” spectrum bands. In this masterplan, the ISM band of 2.4 GHz has been given a low review priority, which means that it will not undergo changes for the next 4 years at least. This is important for designers as it ensures usability within this time frame.

Second, utilisation of the spectrum and the amount of ambient power that is available are important factors for considerations in selecting the band for RF EH designs. Among the five spectrums allocated for wireless services, specifically for indoor environment, 2.4 GHz ISM band under SRD is most widely used [46]. Examining the amount of power available, the table “Technical Requirements for SRD” in Annex 2 [45] lists among others, the limit of RF output power for each SRD applications. From this table, wireless video transmitter and other SRD applications have less than 100 mW EIRP, Bluetooth applications have less than 100 mW EIRP, and lastly, wireless local area network (WLAN) has less than 200 mW EIRP.

Similarly, for wider area communication services, the Spectrum Management Handbook [47] shows the various communications services and their allowed maximum power levels. It states the power ranges from 100 W for mobile services, 50 W for paging service, and 25 W for trunk radio. Among the three services, mobile services has the most number of users. Therefore, to harvest RF energy outdoors, mobile services is the most suitable choice due to the number of users and the maximum transmitter power. From the same table, there is also an overlap between 2G and 3/4G bands. However, as 2G has ceased operations from 12th April 2017 [48] as announced in 5th June 2015 in Singapore, only 3G and 4G bands are available. Generally, the frequency bands to consider spans from 882 MHz to 2025 MHz, and 2500 MHz to 2680 MHz. Therefore, these two most utilised applications are the most suitable for RF energy harvesting.

To determine which band is more suitable for indoor RF EH in a building, a comparison of the amount of energy available indoors is done. The typical distance from a mobile base station is between 30 m to 50 m from the base of the antenna as given in [49]. This translates to a maximum power density of 0.021 W/m^2 to 0.13 W/m^2 for 2G network depending on the height, assuming cell sectoring with 21 channels operating simultaneously. The survey from [50] also shows that the maximum power density is about 0.06 W/m^2 .

However, moving away from 2G, the power density for 3G in the outdoors is lower at 0.0024 W/m^2 . When indoors, due to attenuation and reflections from the façade, the power density is

much lower. According to empirical studies done in [51], the average measured outdoors and indoors received power is about -45 dBm and -56 dBm respectively. The distance of the outdoor measurement is taken 120 m from the base station, while the indoor measurements were taken about 7 m from the façade on average. It is difficult to compare on the same basis for the studies done in [50] and [51] using the same units as information regarding the transmitters such as type of antennas deployed at the base station is not known, and thus the conversion between W/m^2 and dBm is not possible. However, this still gives us an idea about the amount of power available comparing indoors, outdoors and between 2G and 3G.

To compare this with the 2.4 GHz ISM band, we will only look at WiFi as WiFi is predominantly more pervasive and has higher power output than Bluetooth and Zigbee in buildings. Due to lack of survey of indoor WiFi power density, partly due to the different deployment of wireless routers in different settings, there is lacking literature that directly presents the indoor WiFi power density. To establish an understanding of the 2.4 GHz WiFi power density, calculation is done to give an estimation.

Friis transmission equation is given as [52],

$$\frac{P_r}{P_t} = \left(\frac{\lambda}{4\pi R} \right)^2 G_t G_r \quad (2.1)$$

where P_t , P_r , are the transmitter power (W) and receiver power (W) respectively, λ is the wavelength (m) of the incident wave, R is the distance (m), and G_t , G_r , are the gain of the transmitter and receiver antenna.

Converting (2.1) to a logarithmic form with unity gain for both antennas, the received power is written as

$$P_r = P_t - 20 \log f - 20 \log R + 147.55 \quad (2.2)$$

With a typical range of a WiFi router of 20 m, and assuming both the antennas with maximum directivity point towards each other of the same polarisation, the received power is -43 dBm. Comparing this with the amount of power received in [51] (-45 dBm to -56 dBm), it can be seen that the amount of WiFi RF power that can be harvested from a single router is higher than a single mobile base station at a distance of 120 m. In a typical office setting where there is more than one WiFi router, and depending on the number of users, the number of routers can be more, and hence amount of power available will be more than -43 dBm. Therefore, the amount of power available indoors from 2.4 GHz WiFi is greater than the amount of power available from mobile stations indoors.

In addition, in November 2015, IEEE ComSoc has an infographic that summarizes the power of WiFi [53]. It shows that 71% of all mobile communications flow through WiFi, and 60% of carrier network traffic will be offloaded to WiFi by 2017, with more than 7 billion WiFi enabled devices.

Lastly, with Verizon, Qualcomm, Alcatel-Lucent, Ericson, and Samsung promoting LTE-U, or LTE unlicensed in the 2.4 GHz spectrum, this ISM band will be increasingly crowded, creating greater opportunities for RF EH in this frequency range.

2.2 RF Energy Harvesting Antenna

It is known that the amount of energy that can be harvested from ambient RF is limited. From studies done in [54]–[57], they show that among the many available sources such as mechanical vibration, thermal, light (solar or artificial) and ambient RF, ambient RF remains the lowest power density available for harvesting. Although having the lowest power density available for harvesting, ambient RF is typically available at all times. This is very suitable for small power devices such as indoor sensors as they do not require high power but can benefit from a continuous supply of small power. This form of power may even allow devices to be designed without any energy storage.

In a RF EH system, the architecture typically consists of an antenna, followed by a matching circuit, and a rectifier/voltage multiplier [58][59][64] before going to the load or storage as shown in Fig. 2-1.

Fig. 2-1 Architecture of RF harvesting system

Among the stages shown, the antenna and matching circuit are vital stages to the amount of energy that can be collected for rectification. Research done in [60]–[79] show the types of antennas that are used for RF EH in recent years. To compare antennas, figures of merit such as antenna noise temperature, and G/T ratio [80], beam width, polarisation, aperture efficiency, quality factor, and bandwidth [52] can be studied. For energy harvesting antennas, some of the other parameters of interest includes antenna size, gain and polarization [81]. In addition, the antenna and system efficiencies will also directly determine the amount of RF energy that can be harvested.

From this survey, it is observed that many of the recent effort have focused on 2.4 GHz ISM band, 5.8 GHz WiFi band, and mobile communications band. Table 2-1 lists the EH antennas according to their size, operating frequency ranges, S_{11} and gain. For most of the antennas, a minimum S_{11} of -10 dB [60][70] is achieved, with some S_{11} as low as -40 dB [75]. Comparing the gain, for electrically large antennas, the gain can go as high as 9 dBi with the largest dimension 1.53 times the wavelength [78]. For electrically small antennas, the gain is as low as -20 dBi with maximum dimension of one hundredth the wavelength [63]. The higher gain observed for electrically larger antennas, and lower gain for smaller antennas are not unexpected as the gain of the antenna is proportional to its area according to antenna gain equation shown in (2.3) [52].

Table 2-1 Summary of energy harvesting antenna

Reference	Size (mm)	Frequency Range (GHz)	S ₁₁ (dB)	Gain (dBi)
[60]	41 x 44 x 1.778	0.9075	-15	1.2
		2.396	-21.2	1.63
		3.219	-10.3	0.6
		4.883	-11.4	1.24
[61]	70 x 70 x 13.2	1.8 – 2.5	-25	0.35 – 1.85
[62]	92 x 100 x 2	1 – 2	-14.49	N/A
	100 x 100 x 2		-14.81	N/A
[63]	π x 6 x 1.584	0.910 – 0.928	-20	-20.20
[64]	30 x 30 x 10	0.902 – 0.928	-28	1
[65]	33.5 x 17.0 x 0.175	2.2 – 12.1	-35	-3
[66]	61.5 x 48.46 x 0.0254	0.915	N/A	-3
		2.4 ISM	N/A	6
[67]	35.25 x 34 x 0.5	868 MHz	N/A	0.659
[68]	76 x 76 x 1	2.4 ISM	-15	4.9
[69]	152 x 43.2 x 3.5	0.8	-17	-6

Reference	Size (mm)	Frequency Range (GHz)	S ₁₁ (dB)	Gain (dBi)
[70]	30 x 30 x 1.8	0.934 and 1.55	−10.7 and −12.4	N/A
[71]	34 x 34 x 7	2.45	N/A	7.7
[72]	80 x 87 x 20.635	2.45	−25	8.61
[73]	60 x 60 x 30	0.915 and 2.45	−25 and −20	1.87 and 4.18
[74]	62 x 62 x 0.254	0.9	−20	1.8
[75]	38.1 x 38.1 x 1.6	5.8 and 6.1	−40 and −30	7
[76]	110 x 100 x 4.799	2.45	−18.7	5.7
[77]	4.2 x 18.6 x 0.508	5.8	N/A	N/A
[78]	79 x 21 x 0.7572	5.8	−30	9
[79]	137 x 137 x 1.6	0.89 – 0.96	−19	8.5

$$\begin{aligned}
G &= \varepsilon_{cd} D \\
&= \varepsilon_{cd} \frac{4\pi A_e}{\lambda^2} \\
&= \varepsilon_{cd} \frac{4\pi \varepsilon_{ap} A_{phy}}{\lambda^2} \\
&= \frac{4\pi \varepsilon_{cd} \varepsilon_{ap}}{\lambda^2} A_{phy}
\end{aligned} \tag{2.3}$$

where ε_{cd} is the conduction and dielectric efficiency or radiation efficiency, ε_{ap} is the aperture efficiency, D is the directivity, A_e is the effective antenna area, A_{phy} is the physical antenna area, and λ is the wavelength.

Consistent with (2.3), and can be observed from [60] to [79], antennas that have higher gain such as those in [66],[71],[72],[75],[78], and [79] have maximum dimension greater than the wavelength. Similarly, those electrically small antenna such as [63] which has a small area, the gain is comparatively much smaller.

From this comparison of individual antennas for RF EH, it is noted that the performance of the antennas in terms of S_{11} and gain varies with size and design. Designs that are three dimensional such as those in [64][65][73], multi-layer such as those in [61][75][79], or electrically large like those in [66][72][76][78] have better performance in terms of S_{11} and gain. Those antenna designs that are planar [60][62][69][70] or electrically small [63] typically have comparatively poorer performance. Although planar and electrically small antennas have comparatively poorer performance in terms of S_{11} and gain, they have advantages in other areas such as space and contour conformance. For many practical applications in areas such as consumer, buildings and healthcare electronics, these antennas are crucial in enabling wireless solutions.

Observing those antennas with better performances listed in Table 2-1, it is noted that the manufacturing process and cost of fabrication for three dimensional and multi-layer designs may be prohibitive for mass production, and thereby reducing adoption and deployment in buildings. For RF EH in buildings, as the amount of power that can be harvested is low, it is therefore advantageous to have an antenna that is low cost, scalable for different power requirements and surfaces, and easy to fabricate and deploy.

In the context of buildings, especially when there are abundant large surfaces such as the façade, wall partitions, ceilings, and floors, the planar area for the antenna is flexible enough to accommodate different antenna sizes depending on the amount of harvested power required. Antenna array, which has this design flexibility is very suitable for such application. As shown

from Table 2 in [83], the gain of the antenna array, and thus the amount of energy that can be harvested in a given direction increases as the number of elements increases together with the area.

With the application of RF EH in buildings and considering the advantages of antenna array such as flexible array sizes, conformal possibilities, and potentially lower fabrication costs, it is therefore interesting to examine RF EH antenna arrays.

2.3 RF Energy Harvesting Antenna Array

Conventionally, antenna arrays have been used for beam steering [84]–[86], null steering [87][88], shaped beams [89]–[91], and to obtain higher directivity/gain [92]–[94]. In any of the applications, the design of antenna array is affected by at least 5 factors [52]. They are listed as follows.

- The geometrical configuration of the overall array
- The relative displacement of the individual elements
- The excitation amplitude of the individual elements
- The excitation phase of the individual elements
- The relative pattern of the individual elements

The resulting far-zone E field from an array of identical elements is given by

$$E(\text{total}) = [E(\text{Single element at a single reference point}) \times [\text{array factor}]] \quad (2.4)$$

From (2.4), which is also known as the Principle of Pattern Multiplication, it can be seen that if each element of the array is the same, then the array factor, AF is the main contributor of the resulting far-zone E field. In the case of N-element linear array of uniform amplitude and spacing, AF is written as

$$AF = \sum_{n=1}^N e^{j(n-1)(kd \cos \theta + \beta)} \quad (2.5)$$

where k is the wave vector of $\frac{2\pi}{\lambda}$ for a plane wave, d is the element spacing, θ is the arrival angle of a wave relative to the array, and β is the difference in the phase excitation between the elements.

In addition, from (2.5), other than individual elements, d and β are the two main factors that affect AF. For the case of energy harvesting antenna array, it is desirable to have an array that occupies a large area. However, when this is achieved by increasing the distance between each element, the negative result is the generation of grating lobes. These grating lobes are not desirable in some cases as it reduces the beam width of the main lobe, which in turn reduces the gain [95].

Considering this, for broadside arrays, the largest spacing between the elements is thus less than one wavelength, while for end-fire arrays, the maximum spacing is less than one-half of the wavelength [52]. This will need to be balanced with the number of elements that are present in the array. Typically for microstrip patch antenna arrays, the element spacing is chosen to be less than one free space wavelength to avoid grating lobes, but more than one-half wavelength to provide enough space for feed lines and impedance matching patterns, to achieve higher gains and reduce coupling effects [115].

As an illustration, using isotropic antenna array, grating lobes caused by spacing more than the wavelength is shown in Fig. 19.3.2 in [82]. As the element spacing increases from twice the wavelength to eight times the wavelength, the number of grating lobes increases. The number of grating lobes increased from six (for a d of 2λ), to fourteen (for a d of 4λ), to thirty (for a d of 8λ). This increase comes at a cost to the beam width of the main lobe.

On the other hand, when the array elements are placed too close to each other, the effects due to mutual coupling between the elements cannot be ignored. Effects such as changing the radiation pattern, array manifold, and matching characteristics of the elements can occur [96]. In addition, the gain of the array may also be affected when the elements are placed too close together [97].

In the past, for many RF EH antenna arrays, the distance between elements cannot be completely designed based on the coupling effects or gain of the array considerations. The reason is due to the allocation of space for impedance transformers that are needed for matching different elements [98]-[106]. Each antenna array element is interconnected by impedance transformers typically either in the form of corporate feed network, or inline series feed network. Either method requires space between array elements, which in turn reduces the number of element separation possibilities during design. However, in some cases, such as tuning the DC power pattern, the mutual coupling can be used to influence the design [107]. In addition, to reduce grating lobes, having the elements closer together to vary the mutual impedances of the elements can also be useful [108].

With the aim of using the element separation as a parameter for designing the array, it is important to identify the type of RF EH array topology that is suitable. In the literature there are commonly two different types of RF EH topologies. The first topology (see Fig. 2-2(a)), adopts the conventional method of antenna array design. Individual elements are combined through array feed network, which then goes through a RF-DC converter to rectify the combined RF into DC. The second method, on the other hand, individually convert RF to DC at each element, and then combine the resulting DC (see Fig. 2-2(b)) before sending to the load. Due to the nature of DC combiner, spacing occupying impedance transformers can be removed between the elements, or thinner traces can be used. This second method makes it possible to have a greater flexibility in adjusting the element separation to change the mutual coupling between the elements to some advantage [109]-[111]. Due to this, recently more RF EH work has use this topology successfully [112]-[114].

Fig. 2-2 Block diagram of RF to DC topologies (a) RF combiner topology (b) DC combiner topology

Most of the work done in the past for RF EH antenna array has been on microstrip or patch antenna elements. This means that the size of each element approximately follows the dimension of the conventional design formulas as shown in (2.6) [115].

$$L = 0.5 \frac{\lambda}{\sqrt{\epsilon_r}} - 2\Delta L \quad (2.6)$$

where L is length of the patch (in the plane of the feed line), λ is the wavelength, ϵ_r is the relative permittivity of the dielectric, and ΔL is the fringing length as shown in (2.7) [115].

$$\Delta L = 0.412 \frac{(\epsilon_{re} + 0.3) \left(\frac{W}{t} + 0.264 \right)}{(\epsilon_{re} - 0.258) \left(\frac{W}{t} + 0.8 \right)} t \quad (2.7)$$

where W is width of the patch which is used to adjust the input impedance, t is the thickness of the substrate less than 0.02λ , ϵ_{re} is the effective dielectric constant given by (2.8) [115].

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{10t}{W}\right)^{-0.5} \quad (2.8)$$

For example, if high efficiency is required, the formula for W can be as shown in (2.9) [115].

$$W = \frac{\lambda}{2} \left(\frac{\epsilon_r + 1}{2}\right)^{-0.5} \quad (2.9)$$

With these formulas, the size of each element is about 0.25λ . Table 2-2 shows a summary of the size of the elements used in the RF EH arrays. From this table, it is observed that the area utilisation of the antenna array is not high due to the large element spacing.

Table 2-2 Dimensions of elements and spacing for RF EH antenna arrays

Reference	Element size	Element spacing
	(fraction of λ)	
[98]	0.35λ	0.587λ
[99]	0.224λ	0.32025λ
[100]	0.46λ	0.59λ
[101]	0.389λ	0.5255λ
[102]	0.432λ	0.648λ
[103]	$\sim 0.2\lambda$	$\sim 0.6\lambda$
[104]	$\sim 0.19 \lambda$	0.333λ
[105]	Not stated	Not stated
[106]	0.33λ	Not stated

To have a higher area utilisation, the second method where individual RF to DC rectifier is used for each element is needed. This is made possible by the advancing technology in RF components that allows them to be manufactured in smaller packages, with lower energy requirements, and at a decreasing cost. The common rectifying elements used are SMS7630-79 Schottky diode, SMS-2820 Schottky diode and SMS2860 Schottky diode just to name a few [81]. Thus, in recent years, RF EH is starting to go in the direction of packing more elements in an antenna array, each with their own rectifier. This reduces the use of impedance transformers, and each element can be considered an energy source, very much like solar cells. Table 2-3 lists some of the recent efforts to harvest RF energy using array of rectenna and DC combiner topology. Especially seen from work done in [112] and [113], the element spacing is reduced significantly, and the area utilisation is increased significantly compared to those in [98]-[106].

Table 2-3 Dimensions of elements and spacing for rectenna array

Reference	Element size	Element spacing
	(fraction of λ)	
[112]	0.153λ to 1.38λ	Not stated. Very close
[113]	$\sim 0.253\lambda$	0.155λ
[114]	0.164λ	$\sim 0.8\lambda$

From RF EH efforts that went from single antenna of different designs, to planar microstrip patch antenna arrays, and then to rectenna arrays, the trend of RF EH is moving towards having many small rectennas packed as close together as possible to increase the amount of DC power that can be converted.

Following this popular research direction of packing more elements in an array and decreasing the element separation to increase the area utilisation, the design of the array approaches the concepts used in metamaterials, MM.

2.4 RF Energy Harvesting Array Element Size Inspired by Metamaterial

Metamaterial is defined as “artificial effectively homogeneous electromagnetic structures with unusual properties not readily available in nature” [116]. To explain further, “an effectively homogenous structure is a structure whose structural average cell size p is much smaller than a quarter of wavelength, $p < \frac{\lambda_g}{4}$ ” [116]. This means that metamaterials are materials made up of usual materials such as copper and plastics, arranged in a specific manner that can influence the behaviour of EM waves. The construction of the cells in terms of geometry, orientation, size, and shape, when placed together in a certain manner, create a new type of material that cannot be found in nature.

This work is first made known in 1968 by V.G. Veselago in his paper “The electrodynamics of substances with simultaneously negative values of ϵ and μ ” [117]. In this paper, he stated that for an isotropic substance, the dispersion equation can be simplified to (2.10)

$$k^2 = \frac{\omega^2}{c^2} n^2 \quad (2.10)$$

where k is the wave vector of a monochromatic wave, ω is the frequency, c is the speed of light in vacuum, and n is the refractive index of the substance given by (2.11) [117].

$$n = \pm \sqrt{\epsilon_r \mu_r} \quad (2.11)$$

where ϵ_r is the relative permittivity, and μ_r is the relative permeability.

A point to note is the original use of dielectric constant versus the later use of relative permittivity. There is some difference between the two, however, for the scope of this thesis, relative permittivity is considered.

In 1996, research in [120][121] show the requirements of the cell spacing in order to design a metamaterial.

According to classical formula for plasma frequency, ω_p as shown in [121],

$$\omega_p^2 = \frac{2\pi c_0^2}{a^2 \ln\left(\frac{a}{r}\right)} \quad (2.12)$$

where c_0 is the speed of light in vacuum, and a and r are the sizes of the lattice and radius of the wire respectively.

If the wires are not thin compared to the lattice size, then $\ln\left(\frac{a}{r}\right) \approx 1$ according to [120], and the free space wavelength at the plasma frequency, λ_{0p} becomes (2.13) as shown in [121],

$$\lambda_{0p} = 2\pi \frac{c_0}{\omega_p} \approx a\sqrt{2\pi} \quad (2.13)$$

From (2.13), and based on the calculation of penetration depth assuming at half the plasma frequency, the significance of this is that if the radius of the wire is not small compared to the lattice size, then radiation does not penetrate the lattice well. Thus, for a structure to behave as a metamaterial, the wires are required to be thin compared to the size of the lattice.

In the following year, Sir Pendry suggested that for unit cells in a metamaterial to act like atoms in a matter, the size of each unit cell, a has to be as shown in (2.14) [122].

$$a \ll \lambda = 2\pi c_0 \omega^{-1} \quad (2.14)$$

If this is to be adapted to the application of rectenna array, the ratio of the element size to the radius of the wire needs to be determined. Assuming the plasma frequency is equal to the

resonance frequency of the array elements, from (2.12) and (2.14), the size of each element, a can be written as

$$a < \frac{\lambda}{\sqrt{2\pi \ln\left(\frac{a}{r}\right)}} \quad (2.15)$$

From (2.15), and plotted in Fig. 2-3, it can be observed that for the ratio of the element size to radius of the wire of up to 64, the size of each unit element is 0.2λ (see Fig. 2-3(a)). At a ratio of 250, the size of each unit element is 0.169λ . Decreasing the size of unit element further to 0.1λ , the ratio of element to wire size increases to more than 8 million (see Fig. 2-3(b)). From this calculation, it is reasonable to set the element size to between 0.1λ to 0.2λ when used in rectenna array design. This signifies that when each antenna array element is limited to a maximum dimension of 0.1λ , it can be treated as a unit cell in a metamaterial. This limitation is appropriate given the current manufacturing capabilities. Even when nanowires are used, where the length to width ratio is greater than 1000, this limitation satisfies the requirement stated in [121]. In the context of antenna array for RF EH in green buildings, this limitation is sufficient as the aim is to use standard PCB technologies which has lower cost compared to other state-of-the-art technologies such as microelectromechanical systems or MEMs.

According to the work done by Sir Pendry, the size of the element is also the size of the lattice [121]. As the lattices are arranged in a grid-like manner, there is no notion of unit cell separation. In antenna array, however, array element separation is a parameter that affects the properties and performance of the array. Depending on the type of array, the element separation ranges from 0.6λ to 0.32λ (see Table 2-2) and 0.8λ to 0.155λ (see Table 2-3). One of the inspirations of designing antenna array using concepts of MM is treating the elements as unit cells, which in turn emulate atoms in matter. Therefore, in this perspective, if the concepts of MM are successfully implemented, the array element separation becomes an additional parameter that can help in the optimisation of the array design.

(a)

(b)

Fig. 2-3 Fraction of wavelength vs ratio of element size and wire (a) up to ratio of 128 (b) up to ratio of more than 8 million

2.5 Electrically Small Antenna

Thus far, the trend in RF EH is towards rectenna array with decreasing element size and element separation distance. Adopting concepts from MM, a suitable range for the size of each element is determined to be between 0.1λ to 0.2λ . With the size of this dimension, and if a requirement of having the element function as an antenna on its own, each of this element can be regarded

as electrically small antenna, ESA. Therefore, in this section, a survey of some ESAs is conducted, and a suitable definition is selected.

According to Wheeler in 1947 [123], and subsequently in 1959 [124], and by others in [125] and [126], the size of the antenna is considered electrically small when

$$ka < 1 \Rightarrow a < \frac{\lambda}{2\pi} \quad (2.16)$$

where a is the radius of the radian sphere.

In the explanation in [125], the interpretation for electrically small can also be taken as $\lambda/2\pi$. In other references, the definition of electrically small is generally taken as any maximum length of the antenna smaller than 0.1λ [127]–[129]. From these definitions, there is a range of values that an antenna can take to be considered electrically small. In summary, if the maximum dimension of an antenna is less than λ/π to 0.1λ , it is considered small.

Table 2-4 lists some of the antennas that are considered electrically small by the authors. It is observed that the dimensions of the ESAs ranges from 0.04138λ to 1.8λ . For lower frequencies, having a dimension following $ka < 1$ is not difficult as the smallest size ESA in the table comes from an operating frequency of megahertz range [132][134][136][137]. As the frequency increases into the gigahertz range, it is more difficult to decrease the size due to limitations in fabrication processes. With this constraint, the smallest size among the gigahertz range ESA is 0.09324λ [131], and this is achieved by loading less than $\lambda/12$ if the ESA with an inductor to lower the resonant frequency. If external loading is not applied, the smallest size of the ESA is 0.1167λ [130] at gigahertz frequency.

To satisfy the 0.1λ requirement of a unit cell in metamaterials, traditional ESA without any inductor loading is very challenging if not impossible at gigahertz frequencies. Even with external passive loading, the components themselves also have to be electrically small, which

have values difficult to find at increasingly smaller packages. The common packages of multilayer chip components are 0603 ($1.6 \text{ mm} \times 0.8 \text{ mm}$), 0402 ($1.0 \text{ mm} \times 0.5 \text{ mm}$), and 0201 ($0.6 \text{ mm} \times 0.3 \text{ mm}$) sizes. As the size decreases, the choice of values also reduces.

Metamaterial, MM inspired antennas is one of the concepts that can be investigated to modify the performance of ESA. Since the 1990s, MM has been implemented successfully into many applications [141]–[144]. These applications include using MM in optics, superlens, creating artificial magnetics and dielectrics, cloaking, and using it in transmission lines. Among them, the work done by a team led by Prof Richard W. Ziolkowski closely relate to using MM in ESA designs. His work is summarised in [144] which includes research efforts that show the fundamentals of MM ESA [145][146]. He established the definition of an ESA as one that has the radius of the radiansphere less than $\lambda/6$, or less than $\lambda/12$ if ground plane is present (as only half a radiansphere is involved). This means that an antenna is considered electrically small if the maximum dimension is less than 0.333λ , or 0.1667λ if ground is present.

When incorporating MM with ESA, an important concept that is taken from MM is the concept of resonance. As proposed by Sir Pendry, and one of the most widely used MM unit cell structure is the split ring resonator, SRR. This structure is important to MM due to the operating principle that at the desired operating frequency, the ring and the gap between the ring form a LC resonating circuit [147]. When this concept of SRR is incorporated into the design of ESA, the ESA using SRR can form a unit element of an array that can resonate at the desired operating frequency. This type of antenna has been researched in recent years, and they are termed metaresonator antennas [149]–[155] as opposed to other types of antennas such as composite right/left hand based antennas, miniature antennas based on MM loadings, and antennas loaded with metasurfaces [148].

In addition to the seven examples given in [112], there are also some other examples that use the principles of SRR or CSRR and apply them to antennas [156]–[159].

Table 2-4 Summary of dimensions of ESA

Design	Fraction of wavelength	Reference
Three dimensional	0.1167λ @ 2.4 GHz	[130]
Passive component loading	0.09324λ @ 2.4 GHz	[131]
Spiral	0.063λ @ 700 MHz	[132]
Multi-layer	0.252λ @ 1.1 GHz	[133]
Helical	0.08λ @ 408 MHz	[134]
Three dimensional	0.106λ @ 482 MHz	[134]
Three dimensional (PIFA)	1.8λ @ 49 GHz	[135]
Three dimensional	0.0557λ @ 300 MHz	[136]
Folded Spherical Helix	0.04186λ @ 299.8 MHz	[137]
Cylindrical Folded Helix	0.5936λ @ 301.1 MHz	
Matched Disk Loaded Dipole	0.05809λ @ 294.4 MHz	
Matched Spherical Cap Dipole	0.04138λ @ 296.3 MHz	
Matched Spherical Resonator	0.09899λ @ 298.2 MHz	
4-Arm Folded Spherical Helix	0.09899λ @ 298.7 MHz	
5-Arm Folded Spherical Helix	0.10345λ @ 312 MHz	
6-Arm Folded Spherical Helix	0.10345λ @ 312 MHz	
Planar	0.1349 to 0.1951λ @ 1.9 GHz	[138]
Planar	0.2448λ @ 2.4 GHz	[139]
Planar	0.165λ @ 2.4 GHz	[140]

From those designs [149]–[158] listed in Table 2-5, it is observed that the smallest size from this group of metaresonators is 0.1λ at gigahertz frequencies. This is about 10% reduction in size compared to the smallest ESA at such frequencies listed in Table 2-4.

Table 2-5 Summary of dimensions of metaresonators

Design	Fraction of wavelength	Reference
Planar	0.158λ @ 3.62 GHz	[149]
Multi-layer	0.1λ @ 3.1 GHz	[150]
Planar	0.0427 @ 305 MHz	[151]
Three dimensional	0.0828λ @ 403 MHz	[152]
Planar	0.123λ @ 41 GHz	[153]
Planar	0.2913λ @ 4.37 GHz	[154]
Planar	0.198 @ 5.8 GHz	[155]
Planar	0.197λ @ 1.55 GHz	[156]
Planar	1.556λ @ 3.0 GHz	[157][159]
Planar	0.589λ @ 4.245 GHz	[158]

From the previous Section 0, it is determined that the element size is between 0.1λ to 0.2λ . In this section, it is observed that metaresonators have a range of sizes of more than 0.1λ . Bringing together the review results obtained from these two sections, it can be stated that to combine the concepts of MM with RF EH antenna array, each element can be considered as an ESA with a size of 0.1λ . By constraining the maximum size of the ESA to 0.1λ , the strictest definition of ESA is applied. Using this definition, the antenna design can be considered electrically small, and when used as an array element, it can satisfy the definition of a unit cell in metamaterials.

2.6 Metamaterial Rectenna Array for RF Energy Harvesting

In recent years, there have been successful attempts to harvest RF energy using metamaterial inspired antennas. Incorporating MM into antennas, it enables the antenna to be electrically larger than its physical size. In this sense, much of the effort in harvesting RF energy have yield good results [160]–[164], with one design having more than 80% RF to DC efficiency at 3

dBm [160]. The overall system efficiency would not be investigated further due to the scope of this thesis. The focus here is to investigate the use of electrically small metaresonator as a RF EH array element rather than designing a RF EH system.

To develop an understanding of the potential of using MM antenna array for RF EH, it is useful to understand the efficiency of solar cells. From Shockley–Queisser limit, the maximum efficiency is 33.3% [165] for cells with a single p-n junction, and a theoretical maximum of 86% [166]. Contrasting with research done in [167] and [168], there is good potential for MM inspired array to harvest RF energy in terms of efficiency when compared to solar cells.

Table 2-6 summarizes the MM inspired arrays design and performance differences. Both rectenna arrays come in planar form and they are printed using PCB technologies with [167] using Rogers Duroid RT5880 while [168] uses FR-4 substrate. The element sizes for both designs are relatively similar at 0.161λ and 0.202λ respectively. An observation for these two designs is the disconnectedness of the elements. In [167], each of the element is connected to its individual resistor, while in [168], each element is connected to a diode. The significant differences in the RF to DC efficiency is also due to the different power levels that are used, and how efficiency is regarded. There are several ways the arrays or individual elements can be connected to the load or connected to the rectifying circuit. The different methods are not discussed further as the focus is on the antenna and not the entire system.

Table 2-6 Summary of RF energy harvesting rectenna array

	Square shape SRR antenna [167]	I shape MM antenna [168]
Substrate	Rogers Duroid RT5880	FR-4
Frequency	5.8 GHz	2.18 GHz
Element size	0.161λ	0.202λ
RF to DC Efficiency	40% to 75%	96.2%

From these reviews, an understanding of the trend of RF EH is developed. In addition, having understood the advances in DC combining antenna array for RF EH, MM, and ESA, it is learnt that there exists a possibility of adopting metaresonator based ESA into a RF EH antenna array. Thus, the design of an array using such a design operating within the 2.4 GHz WiFi range will be demonstrated in this thesis.

2.7 Metaresonator Arrays in Building Materials

An important observation obtained from this survey is that both resonator-based antenna, and ESA have limited bandwidth. In addition, when the concept of resonance such as addition of SRR is used to improve the matching of ESA, the bandwidths of such antennas are also narrow [169]. To improve the performance of ESA, there have been efforts to determine the bandwidth limits [198], efficiency limits [178], radiation limits [170], and also ways to increase the bandwidth. The ways of increasing the bandwidth include resistive loading [171], optimising length and breadth ratio [172][173], geometry [174], and dielectric modifications [175]. As the context for this thesis is aimed at RF EH in the built environment, specifically for indoor spaces, the focus would be on the investigations of using the environment to improve the performance of the antennas. One of the identified application is the embedding of RF EH antennas in building materials to increase the bandwidth and the related effects on the performance of the embedded antenna.

Several techniques from past research efforts in miniaturising antenna are described in [176]. Among the many methods, designing antennas with engineered substrate is considered one of the techniques. Typically, the substrate of the antenna is modified to affect the antenna in certain ways. One of the method in this technique is the synthesis of material with metallic particles mixed with the dielectric. However, as pointed out in [176], more investigation needs to be carried out before this technique is widely accepted. In this thesis, this is done with metallic and/or magnetic particles to investigate the effects this technique has on the embedded metaresonator array and the EM property of the resulting material.

As we already know now that wireless communications are essential in buildings for many reasons, including internet of things (IOT) applications. However, the materials used in buildings do not enable good propagation of wireless signals within, or in and out of buildings. This is a practical challenge for antennas and especially for embedded antennas in building materials. With rising interest to use such embedded antennas in building materials for concrete health monitoring [180][181], wireless powering of embedded sensors [183] and radio-frequency identification, RFID applications [184], studies have been conducted to investigate the effects of building materials on electromagnetic, EM propagation.

It is known that common building materials, such as concrete, gypsum and plaster, do not allow good EM propagation due to their high shielding effectiveness, SE. Direct measurements [185][186], and analysis using finite-difference time-domain, FDTD [187], and method of moments, MOM [188] techniques show that building materials have more than 80 dB to 100 dB of SE.

One way to improve the performance of concrete is by adding magnetic and/or conductive particles as admixtures into the cement mix. Previously, the addition of particles in concrete focused on the need to dispose of industry residues (an environmental aspect). Existing studies of this composite material with additions of magnetic and/or conductive particles have been limited to the performance of concrete and its durability [189]–[192]. Employing the use of such particles for the improvement of EM properties of the concrete structure has not been done for embedded antennas to date. Furthermore, the use of conductive concrete for EM applications to date has rather been focusing on improving the SE of concrete.

Some of the past work show that conductive concrete for EM applications using carbon fibers in polymer concretes [193], and electro conductive concrete [194][193] with mortar blocks for frequencies from 30 MHz to 5 GHz improves the SE of materials. The SE achieved are 65 dB at 500 MHz, 75 dB at 1 GHz, and 95 dB at 1.5 GHz. Other researchers also used different concrete mixtures with steel and carbon fibers, and different grades of carbon powder, achieving an SE of 52 dB at 1 GHz [195], or using concrete mixtures with carbon fibers

attaining a SE of approximately 20 dB to 30 dB between the frequencies of 1 GHz and 2 GHz [196].

However, a recent study [210] shows that by combining non-metallic and metallic particles in a material, it creates a phenomenon of “microwave passage” that allows deeper penetration of microwave of 2.45 GHz into the material. While the concentration is to be adjusted to suit the building industry, it is possible to improve the transmission of wireless signals in a material by adding metallic particles.

Therefore, in this thesis, investigations into how some common admixtures added to building materials can improve the EM transmission property of the resulting materials are done. The findings will contribute to the research efforts done regarding the SE of such materials.

Chapter 3

Scalable Electrically Small Metaresonator and Element

3.1 Overview

From the previous chapter, we understand that it is possible to design a RF EH antenna array using concepts of MM. In this chapter, concepts of split ring resonators, SRR are first used to design a 0.1λ planar electrically small metaresonator. This metaresonator can resonate well at the desired frequency without any passive loading elements of resistors, inductors or capacitors. In addition, the geometry of this metaresonator enables scalability, allowing it to tessellate into a larger array as an array element. Next, using this metaresonator element, an array is designed and fabricated. Further analysis of this metaresonator array at the different feed points contributes more knowledge into the behaviour of this type of antenna array.

3.2 Selection of Geometry for Electrically Small Metaresonator

There are three conditions to satisfy when designing this ESA, which can also be used as an array element. First, the size of this antenna is limited to a maximum of 0.1λ to combine the

concepts of MM and ESA into an antenna array. The advantage in addition to the assumption of uniform current distribution, is the application of lumped circuit model analysis [177]. One way to design such a gigahertz ESA is to design it as an electrically small metaresonator.

The second condition that we consider is that this electrically small metaresonator can also be used as an array element that can be tessellated into an array. The elements are also required to pack close together to increase area utilisation and reduce grating lobes.

Lastly, the cost of fabrication is also an important consideration. In the application of this metaresonator in buildings, where the size can potentially be large and quantity numerous, low cost can encourage adoption and implementation. Therefore, we will restrict the design to standard PCB technologies. In addition, conventional low-cost FR-4 is also selected as the material for substrate. A point to note is that in PCB technologies, there are four classes of printed boards under the Association Connecting Electronics Industries (IPC) standards. Class 1 is for general electronic products, Class 2 is for dedicated service electronic product, Class 3 is for high reliability electronic products, and Class 3/A is for space and military avionics. As those names suggest, Class 1 is the cheapest but with the lowest standard, while Class 3/A is the most expensive but has the highest standard. In most cases, Class 2 has a good balance between cost and standard.

Given a size of 0.1λ , the area must be as big as possible to have the highest possible gain according to (2.3). In addition, with the considerations of tessellation of this ESA as an array element, this electrically small metaresonator is chosen to have a shape of a regular polygon. A regular polygon is preferred as it has good reflection and rotational symmetry. With the selection of a regular polygon, the next step to find out which regular polygons can be tessellated.

The condition for tessellation can be written as

$$\sum_{i=1}^m \theta_i = 2\pi \quad (3.1)$$

where θ_i is the interior angle of the vertices, m is the number of vertices at a point.

For a n -sided regular polygon,

$$\begin{aligned} \theta_1 &= \theta_2 = \theta_3 = \dots = \theta_n \\ \therefore m\theta_n &= 2\pi \\ m\frac{\pi}{n}(n-2) &= 2\pi \\ (n-2)(m-2) &= 4 \end{aligned} \quad (3.2)$$

As the integer factors of 4 are 1, 2, and 4,

when $n = 3, m = 6 \Rightarrow$ Equilateral Triangle

when $n = 4, m = 4 \Rightarrow$ Square

when $n = 6, m = 3 \Rightarrow$ Regular Hexagon

$\therefore$ These are the only three regular polygon that can be tessellated.

Among these three polygons, hexagon has the largest area. The formula for the area of a regular hexagon is $\left(\frac{3\sqrt{3}}{8}\right) R^2$, while the areas of square and equilateral triangle are $\left(\frac{1}{2}\right) R^2$ and $\left(\frac{\sqrt{3}}{4}\right) R^2$, respectively. This gives the regular hexagon an area that is 1.3 times that of the equilateral triangle and 1.5 times that of the square. Hence, the geometry of the ESA chosen is of the shape of a hexagon as gain is proportional to its area (2.3).

Using this geometry, the type of ESA to design is determined next. Typically, the two types of small antennas are electric dipole and magnetic loop [178]. From the survey in the previous chapter, it is understood that concepts of resonators such as SRR can be incorporated into the design of ESA. An SRR takes the shape of a loop, with a split to form a LC circuit. To combine this concept into the ESA design, an electric dipole type of antenna is chosen. For this design, one of the six sides of the hexagon is selected to include the split. Another split is also required to accommodate the feed point to feed the ESA. In order to determine a suitable location for this feed point, an understanding of radiation resistance of an ESA is necessary.

Radiation resistance indicates how well an antenna converts energy into electromagnetic radiation. For an ESA, due to its size, the radiation resistance is low. Radiation resistance, R_r for an infinitesimal dipole can be derived as

$$R_r = 80\pi^2 \left(\frac{l}{\lambda}\right)^2 \quad (3.3)$$

where l is the overall length according to [52]. A wire antenna is considered an infinitesimal dipole if $l < \lambda/50$. For small dipoles, the radiation resistance can also be shown [52] to be

$$R_r = 20\pi^2 \left(\frac{l}{\lambda}\right)^2 \quad (3.4)$$

These two equations suggest that one of the way to improve the radiation resistance of an antenna is by increasing the overall length of the dipole.

With the aim of having as long an overall length as possible for the ESA, the feed point is selected to be on the opposite side of the split. A sketch of the ESA shown in Fig. 3-1. This basic concept forms the foundation for further designs. From this sketch, it resembles a single SRR with an addition of a feed point on the opposite side of the split. Having the additional gap to accommodate the feed point, the next section analyses how this may affect the LC

resonant concept of an SRR.

Fig. 3-1 Concept of hexagonal shaped ESA

3.3 Analysis of Split Ring Resonator with Feed Point

From Fig. 3-1, the hexagonal metaresonator consists of a top split and a bottom split. The top split functions as a capacitor while the bottom split acts as an unintentional capacitor meant for the feed point. Despite this, it may be able provide a certain amount of capacitive effects for the metaresonator. This metaresonator can be mathematical modelled using lumped circuit model as shown in Fig. 3-2.

Fig. 3-2 Lumped model of hexagonal metaresonator (a) lumped circuit model (b) simplified model

Based on the simplified model as shown in Fig. 3-2(b), the impedance of the hexagonal metaresonator is

$$\begin{aligned}
 Z &= \left(R + j\omega L + \frac{1}{j\omega C_u} \right) // \frac{1}{j\omega C_b} \\
 &= \left[\frac{j\omega C_u}{(1 - \omega^2 LC_u) + j\omega RC_u} + j\omega C_b \right]^{-1} \\
 &= \frac{\omega RC_u - j(1 - \omega^2 LC_u)}{\omega(C_u + C_b - \omega^2 LC_u C_b) + j\omega^2 RC_u C_b}
 \end{aligned} \tag{3.5}$$

Rearranging Z into real and imaginary parts,

$$Z = \frac{\omega(RC_u^2 + RC_u C_b - RLC_u^2 C_b - \omega RC_u C_b + \omega^3 RLC_u^2 C_b) - j[(1 - \omega^2 LC_u)(C_u + C_b - \omega^2 LC_u C_b) - \omega^2 R^2 C_u^2 C_b]}{\omega^2[C_u + C_b - \omega^2 LC_u C_b] + \omega^4 R^2 C_u^2 C_b^2} \tag{3.6}$$

For resonance, $Im(z) = 0$,

$$[(1 - \omega^2 LC_u)(C_u + C_b - \omega^2 LC_u C_b) - \omega^2 R^2 C_u^2 C_b] = 0 \tag{3.7}$$

According to [122], assuming certain conditions, the capacitance between two parallel sections of metallic strips can be written as

$$C = \frac{\epsilon_0}{\pi} \ln \frac{2c}{d} \tag{3.8}$$

where c is the width of the strips and d is the distance between the strips.

From (3.8), it is observed that the relationship is consistent with the general capacitance relationship of $C = \epsilon A/d$, where A is the overlapping area of the plates. The capacitance, C

decreases as the distance d between the plates increases.

If the design of the ESA accommodates a feed point at one side of the hexagonal ring, and a split for resonance on the other side, the distance of the capacitive split will likely be much smaller than the distance of the split for the feed point. Therefore, the capacitances due to the splits can be taken as

$$C_b \ll C_u \quad (3.9)$$

Taking this assumption, (3.7) can be reduced to

$$(1 - \omega^2 LC_u)(C_u) = 0 \quad (3.10)$$

And the resulting resonance frequency can be shown to be

$$f = \frac{1}{2\pi\sqrt{LC_u}} \quad (3.11)$$

When the effect of the feed point is much less than the split on the other side of the ring, the resonant frequency is consistent with the resonant tank circuit.

3.4 Design of Electrically Small Metaresonator Element

To find out the optimal dimensions for operation in the required frequency, optimisation in simulation software such as CST® Microwave Studio (CST MWS) can be used. Using CST, a model of the hexagonal metaresonator is built as shown in Fig. 3-3.

This model is drawn as close to represent an actual antenna as possible. This includes using limitations in IPC Class 2 PCB technologies, having through-hole vias to bottom layer, and 0402 size component pads (see Fig. 3-4). To facilitate optimisation, the various dimensions are

also modelled as variables.

Fig. 3-3 Single hexagonal SRR (a) front (b) back

Fig. 3-4 Design of via through holes (a) front view (b) perspective view

Table 3-1 summarizes the physical parameters of the ESA that will not vary during optimisation.

To determine the maximum dimension, R of the unit element, the different WiFi channels are identified. There is a total of 14 channels for WiFi 2.4 GHz. However, channel 14 is seldom used worldwide, and thus the highest frequency of the 2.4 GHz WiFi band is taken at channel 13 without any loss in generality. The frequency for channel 13 is 2.472 GHz, and adding the channel bandwidth of 11 MHz, the end of channel frequency is 2.483 GHz. At this frequency,

the corresponding wavelength is 120.8 mm. Therefore, the R in this application is taken to be 12.08 mm (one-tenth the wavelength). This ESA is then optimised and simulated in CST® to study the effects on S_{11} vs frequency of between 2 GHz to 3 GHz with C_u as the varying parameter.

Table 3-1 Summary of physical dimensions

Parameter	Property	Units
Substrate	FR4	-
Substrate thickness	1.6002	mm
Copper thickness	1 (1.4)	oz. (mils)
Minimum trace width	5	mils
Minimum trace separation	3.5	mils
Through hole via separation	75	mils
Through hole via diameter	10	mils
Annular ring diameter	20	mils
Through hole via clearance	5	mils

There are seven different optimization algorithms available in CST® ranging from Classic Powell to Covariance Matrix Adaptation Evolution Strategy (CMA-ES). Classic Powell being a local optimizer is suitable for small parameter variations where the initial value is close to the optimized value, while CMA-ES is a global optimizer capable of searching the optimized value when a close initial estimate value is not available. The disadvantage of global optimizers is that optimization process is computationally expensive, while local optimizer may be trapped in some local minima.

Using CMA-ES optimisation of the top gap width to vary C_u with a limit of 3.5 mils for trace

separation, the lowest point of resonance for this ESA is around 7 GHz (see Fig. 3-5).

Fig. 3-5 S_{11} of hexagonal ESA

At the required WiFi band of between 2.4 GHz to 2.5 GHz, the S_{11} is -2 dB, which is insufficient to function well as an antenna. To function as an antenna, typically the S_{11} is in the range of less than -6 dB for marginal performance and -20 dB or less for good performance. Due to the limitations of PCB technologies of IPC class 2, the trace width used is 3.5 mils, which is the thinnest allowed. Typically, to resonate at a lower frequency, the length of the antenna is increased to increase the inductance. However, the length of this ESA is constrained by its geometry and maximum dimension, and thus, the inductance cannot be increased further.

To lower the resonant frequency, the capacitance can be increased instead according to (3.11). However, even after optimisation, the distance of the gap at the top of the metaresonator reaches the limit of 3.5 mils. This distance is the smallest trace separation allowed under IPC Class 2 and any further increase in capacitance is not possible. Due to these constraints, it can be

concluded that given the limitations of IPC class 2 and substrate material of FR-4, such a planar hexagonal ESA of size 0.1λ cannot be designed to operate at a frequency lower than 7 GHz. To decrease the resonant frequency to operate at the desired 2.4 GHz ISM band, modifying the SRR, and coupling another SRR is considered to improve the matching.

Other approaches are considered to increase either the inductance and/or capacitance without increasing the overall dimension. Extra stubs are added to increase the length of the ESA (see Fig. 3-6(b)), adding an inner SRR (see Fig. 3-6(c)), and adding an inner SRR with stubs (see Fig. 3-6(d)).

Fig. 3-6 Four different approaches (a) hexagonal ESA, (b) with stubs, (c) with SRR, (d) with SRR and stubs

The variables that are used for the optimisation are the length of the stubs, the distance of separation between the inner and outer rings, and the width of the inner and outer traces. The S_{11} results of the four ESAs after optimisations are plotted in the same graph for comparison in Fig. 3-7. From Fig. 3-7, it is noted that the design of Fig. 3-6(d) has the best S_{11} result among the four designs. This study shows that increasing the length of stubs and coupling another SRR improve the S_{11} .

Fig. 3-7 S_{11} comparison of four different metaresonator configurations

Therefore, to further improve the S_{11} , both inner and outer rings are added with stubs. Not only are the rings coupled closely with a minimum allowed trace separation, the stubs are also designed to couple together. The inner ring and the outer rings are placed as close together as allowed under IPC class 2 of 3.5 mils. The variables that are optimised are the length of the stubs and trace width. Fig. 3-8 shows the final model of the electrically small metaresonator after optimisation.

Fig. 3-8 Electrically small metaresonator (a) front (b) back

To compare with the other two regular polygons that can be tessellated as shown in (3.2), additional two models of square and equilateral triangles with the same maximum dimension are created (see Fig. 3-9). The trace width, trace separation, and stubs lengths are optimised. Similar to the hexagonal metaresonator, the trace width and trace separation are both at the limit of 5 mils and 3.5 mils respectively. Therefore, the stubs lengths are optimised for resonance to occur at the required channel 7. Results of the S_{11} are shown in Fig. 3-10.

Fig. 3-9 Other metaresonators (a) square (b) triangle

From the figure of S_{11} comparison among the three metaresonators, it can be seen that all three resonate at the required frequency. It is also clear from this finding that the hexagonal metaresonator has the best S_{11} of less than -25 dB.

Since it is already known that the gain of the antenna is proportional to the area, a simulated comparison of the gain is done on the three metaresonators as shown in Fig. 3-11 for three channels 1, 7, and 13.

Fig. 3-10 Comparison of S_{11} of triangle, square, and hexagon metaresonators

From the graph, the hexagonal metaresonator has the highest gain throughout the band. Although the hexagon shape has 1.3 and 1.5 times the area of the square and triangle shape respectively, the gain however, does not show similar behaviour, even though area and gain are proportional. This is largely due to radiation resistance that is not proportional to the area of the ESA. Although the gain is the highest, it is considered very poor gain. The reason is primarily due to its small size. As shown in [63], when the size is very small, the gain is very

low. In that design, the gain obtained is -20.2 dB. Table 3-2 summarizes the area and gain compared to the hexagonal metaresonator.

Fig. 3-11 Gain comparison between the three metaresonators

Table 3-2 Comparison of area and gain with respect to hexagonal metaresonator

	Area	Channel 1 (2.412 GHz)	Channel 7 (2.412 GHz)	Channel 13 (2.412 GHz)
Triangle	1.5 times	3.66 times	3.57 times	3.37 times
Square	1.3 times	1.04 times	1.07 times	1.13 times

During the optimisation process, it is noted that the length of the stubs can be used to tune the resonating frequency, consistent with (3.11). By varying the length of the stubs, the impedance of the ESA, and thus the matching is changed. Fig. 3-12 shows as the stub length is changed from 0.1 mm to 2.6 mm, the magnitude of S_{11} shifts from the right to the left in a parameter sweep. This shift corresponds to a lowering of the resonant frequency due to increases in

reactance. From the graph, it can also be observed that the optimal stub length is between 0.6 mm to 1.1 mm for the required resonant frequency to be at channel 7 within the 2.4 GHz band.

Thus far, the optimisation tool used for determining the optimal length of the stubs is CMA-ES. This is a global optimiser that can resolve the value of the length of the stub very well, but this process is time and resource consuming.

Fig. 3-12 S_{11} response on varying stub lengths

With the parameter sweep that was done, it can be observed that the optimum length of the stub lies between 0.6 mm to 1.1 mm (see Fig. 3-12). Having this narrower range, a quicker optimisation tool can be used. To verify this, a local optimiser, Classic Powell is used to search for the optimal length of the stubs for values between 0.6 mm to 1.1 mm. After a small fraction of the time, the results show that the optimal length is 0.775 mm. This value is consistent with the value obtained earlier.

The contributions made here thus far show that it is possible to design a planar 2.4 GHz 0.1λ ESA using concepts of SRR from MM to form an electrically small metaresonator.

3.5 Verification of Hexagonal Electrically Small Metaresonator

Fig. 3-13(b) shows the fabricated proposed metaresonator. To validate the performance of the proposed antenna, a S_{11} measurement is done using Agilent EA-6884 Vector Network Analyser (VNA) on the fabricated antenna. A RF coaxial cable is soldered directly onto the 0402 component footprint soldering as shown in Fig. 3-14.

Fig. 3-13 Proposed antenna (a) CST® model (b) fabricated

Fig. 3-14 Back of fabricated metaresonator with RF cable soldered

Fig. 3-15 shows the comparison between the simulated and measured results. The S_{11} of the simulated result shows -8.3 dB, -25.7 dB, and -9.4 dB at channels 1, 7, and 13 respectively, with a -10 dB bandwidth of 2%. For the measured result, the S_{11} is -10.5 dB, -22.9 dB, and -11.5 dB for channels 1, 7, and 13 respectively, with a -10 dB bandwidth of 2.67%. Both results show good agreement with each other at the extreme channels as well as the centre channel. This also shows that the bandwidth of the antenna is wide enough to cover the WiFi band from channels 1 to 13.

Fig. 3-15 S_{11} of fabricated metaresonator compared with simulated results

Fig. 3-16 shows the radiation patterns of the simulated and measured proposed antenna of the two extreme channels at 1 (Fig. 3-16 (a)) and 13 (Fig. 3-16 (c)), as well as the centre frequency at channel 7 (Fig. 3-16 (b)). The radiation patterns of both results show good agreement for all the 3 channels in both the E-plane and the H-plane. In addition, the radiation patterns show consistency when the frequency changes from channel 1 to 13. The patterns displayed for the three channels are also similar to that of a dipole.

Compared to other ESAs (see Table 3-3) the S_{11} is good, and comparable with the other ESAs that are three dimensional or loaded with passive elements. This shows that using the concept of coupling a planar ESA having a size of 0.1λ with SRR, the impedance of the ESA can be well-matched, with S_{11} as good as those antennas that are three dimensional or passively loaded. In terms of gain, while the gain of the electrically small metaresonator is low compared to the other three-dimensional ESAs, the gain is better when compared to a similar planar ESA [63] . This work demonstrates that it is possible to design a planar ESA of 0.1λ in the gigahertz range.

Table 3-3 Comparison of other ESAs with proposed metaresonator

Antenna Design	Size (mm)	S_{11} (dB)	Gain (dBi)	Reference
Planar	0.018λ @ 928 MHz	-20	-20.20	[63]
Three diemnsional	0.1167λ @ 2.4 GHz	-27	2.00	[130]
Loaded with passive component	0.09324λ @ 2.4 GHz	-23	-6.50	[131]
Three dimensional	0.1λ @ 3.1 GHz	-23	0.90	[150]
Planar	0.1λ @ 2.4 GHz	-25	-16.14	This work

One of the objectives of this thesis is to also use this proposed electrically small metaresonator as an antenna array element. The main reason is so that regardless of the scale of the array, from one ESA to multiple elements in an array, this design can be used as a standalone antenna.

As we know scaling an antenna into an array can increase the gain, the next chapter presents an option of increasing the area by tessellating this electrically small metaresonator element into an antenna array. One of the design objectives is to tessellate this element into an array without using impedance transformers.

Fig. 3-16 Measured (solid line) and simulated (dashed line) radiation patterns of proposed antenna (unit: dBi) (a) channel 1, (b) channel 7, (c) channel 13

3.6 Concluding Remarks

Based on the literature review in Chapter 2, it is proven that by limiting the maximum dimension of an array element to be one-tenth of the wavelength, the concepts of MM and antenna array can be combined. To begin designing the array, the concept of the array element is first established. Next, to add flexibility during deployment in a building for WSN, this array element is also designed to function independently as an ESA. However, from the literature, designing a planar ESA of 0.1λ using standard PCB technologies is challenging without passive component loading. This design problem is solved by using concepts of SRR to couple the ESA to form an electrically small metaresonator. The unique feature of this ESA is the addition of stubs for both the ESA and inner SRR. These stubs allow tuning of frequency such that optimisation process can be simplified. The resulting planar ESA is simulated in CST® and verified with a fabricated sample using IPC class 2 processes. The results show that the electrically small metaresonator is well matched at the desired frequency with a S_{11} of between -22 dB to -25 dB. The -10 dB fractional bandwidth as expected of an ESA is between 2% to 2.7%.

The contribution in this chapter shows that a planar ESA of 0.1λ can be designed to operate in the gigahertz range using standard PCB processes without loading with passive components. The proposed ESA is coupled with an inner SRR to improve the matching. Compared with other ESAs, the S_{11} is good while the gain is comparable.

Chapter 4

Scalable Electrically Small Metaresonator Array

4.1 Scalable Metaresonator Array with DC Connection

From the previous section, given the constraints of 1) maximum size of one-tenth the operating wavelength, 2) planar in design, 3) ability to tessellate into an array, and 4) using standard PCB technologies and FR-4 material for the substrate, we determined that the proposed electrically small metaresonator behaves like a LC circuit. Furthermore, it is shown that it is possible to design such a metaresonator to resonate well at the desired gigahertz frequency without using any passive components loading.

In this section, this electrically small metaresonator is used as an element to form an array to increase the gain. Two different types of configurations are designed, and one is selected based on the constraints identified. The selected metaresonator array is then simulated and verified to show that by using this metaresonator element, a metaresonator array can be designed with

each unit element closer to each other, and thus increasing the area utilisation. In addition, due to the way individual elements is designed and interconnected, a study of the performance of this array at different feed points is carried out.

4.2 Selection of Array Configurations

A way to increase the gain of an antenna is to increase the overall area of the antenna as described in (2.3). This can be achieved by designing an antenna with a larger aperture or combining more antennas to form an array. With the design of an antenna shown in Fig. 3-8, it can be used as an element for an antenna array. The first advantage is that this hexagonal geometry selected is suitable for tessellation such that the spacing between each element can be minimal, and the area utilisation is maximised. Next, there can be more than one configuration of array pattern that can be formed by changing the orientation of each element. While many different configurations are possible, not all are suitable with the constraints mentioned. Fig. 4-1 shows two different configurations to explain the suitability of configurations in this application.

Type A configuration in Fig. 4-1(a) shows the elements arranged in a way with the splits facing in the outward direction. This outward direction when interconnected at the back of the substrate, provides a hexagonal way of interconnecting the elements. The CST® software model (see Fig. 4-2) shows how the array looks like from the front and back. Although the connections are neat, the main disadvantage is that this array cannot be uniform when scaled up to a larger size. The interconnections at the back of the array also does not allow easy connections when more elements are added. Despite this, the model is created to demonstrate the feasibility of this configuration as an array.

The second configuration, type B (see Fig. 4-1 (b)), arranges the elements all in the same direction. The CST® model in Fig. 4-3 shows the front and back of this configuration. The interconnect at the back of the array are designed to be vertical and horizontal. Using this arrangement, this array can be scaled to different sizes.

Fig. 4-1 Different configurations of array (a) type A, (b) type B

One of the main advantage of this electrically small metaresonator that has been developed is the ability to function well as a stand-alone antenna. In addition, the length of the stubs can be used for fine tuning when placed in different applications. In this instance, there are two types of configurations of elements to form different arrays. Due to this advantage, all the dimensions such as the trace width, rings separation distance, overall size of element, and via through hole dimensions, except for the length of the stubs can remain constant. This design allows quick tuning and optimisation when the elements are placed into an array.

Fig. 4-2 Type A metaresonator array in CST® (a) Front, (b) Back

Fig. 4-3 Type B metaresonator array in CST® (a) Front, (b) Back

For the optimisation of these two configurations of array, having the stated dimensions pertaining to the ESA remaining constant, only two parameters are needed for optimisation; 1) the length of the stubs, and 2) the distance of separation between each element. With only two parameters to optimise, the optimisation is less computationally intensive. To reduce the computation load further, a sweep using the two parameters to investigate the effects of the two parameters is performed first. This allows a general idea of how the two parameters affect the S_{11} . Next, local CST® optimisation tool similar to the methods used for the single electrically small metaresonator is applied.

The resulting optimised dimensions of the length of the stubs and element separation distance are obtained and listed in Table 4-1.

Table 4-1 Comparison of proposed metaresonator arrays

	Type A configuration	Type B configuration
Element distance of separation (mm)	0.089 (0.000724 λ)	2.032 (0.0165 λ)
Length of stubs (mm)	0.1	0.561

Fig. 4-4 shows the resulting S_{11} plot of type A metaresonator array. Despite having a narrow -10 dB fractional bandwidth of 0.33%, it is able to resonate well at the centre frequency. This shows that this configuration can be matched and resonate well at the centre frequency. However, with the length of the stub at 0.1 mm and an element separation distance of 0.089 mm (~ 3.5 mils), this design will require careful attention during fabrication.

On the other hand, the configuration shown in Fig. 4-1(b) arranges the elements in a different way. The orientation of each element is the same, and thus the direction of the feed points is also the same. This manner of arranging the unit elements allows good scalability as the number of elements increase. To interconnect these elements, the connections at the back of the PCB can also be simplified. As shown in Fig. 4-3(b), the interconnects are suitable for connections even as the size of the array increases.

Fig. 4-4 S_{11} of type A, metaresonator array

Compared to type A, type B has a length of the stubs at 0.561 mm, and element separation distance of 2.032 mm (0.0165λ). The S_{11} plot of type B metaresonator is shown in Fig. 4-5. With the optimised parameters, this metaresonator array is able to resonate well at the centre frequency. At the same time, the S_{11} covering the WiFi spectrum is also less than -10 dB. Compared to type A configuration, the -10 dB fractional bandwidth is wider at 3.07%.

Therefore, for type A configuration, due to the non-uniform orientation tessellating manner the elements are arranged, the non-scalable manner the unit elements are interconnected at the back of the PCB, and the tighter tolerances required during fabrication, it is not a suitable configuration for the proposed application of scaling up unit elements to form bigger arrays. Thus, in terms of scalability, manufacturability, S_{11} and fractional bandwidth, type B metaresonator array is the more suitable configuration.

Fig. 4-5 S_{11} of type B metaresonator array

Although type B is selected as the better configuration compared to type A, both metaresonators

array can be optimised to operate at the desired frequency. This is made possible by the proposed designs used in this chapter. Firstly, the proposed design using stubs to load the top of the ESA element allows changing the length of the stubs to tune the ESA to resonate at the desired frequency. In addition, when these elements are formed into an array, the length of the stubs can also be adjusted to tune the resonant frequency of the array. Secondly, planar unit elements of one-tenth the size of wavelength is realised by combining the concepts of SRR with electrically small antennas to form electrically small metaresonators. This allows each ESA element to function individually as an ESA at the required frequency. Finally, the combination of these concepts also allows the ESA elements to be arranged closer to each other, and thereby increasing the area utilisation.

4.3 Type B Metaresonator Array – Verification and Analysis

4.3.1 Verification

Fig. 4-6 shows the CST® model of type B configuration with the laminate hidden. This view shows the metaresonator array on the top layer of the PCB and the interconnections on the bottom. The top layer and bottom layer are linked with through hole vias modelled using practical dimensions.

To verify the simulated results, this type B metaresonator is fabricated using IPC class 2 processes. The fabricated metaresonator array is shown in Fig. 4-7. The overall dimension of this seven-cell metaresonator array is 40 mm x 40 mm x 1.6 mm. A S_{11} comparison between the fabricated and simulated metaresonator array is shown in Fig. 4-8. The results agree well with each other, showing S_{11} values of less than -25 dB, and a -10 dB fractional bandwidth of 3%.

Fig. 4-6 Type B metaresonator array model in CST® software

(a)

(b)

Fig. 4-7 Fabricated type B metaresonator array (a) Front, (b) Back

Fig. 4-8 S_{11} of type B, metaresonator array

In terms of gain, the gain of the array is also increased at -5.5 dB to -4.7 dB compared to a unit element of -15.4 dB to -17.4 dB (see Fig. 4-9). Table 4-2 lists the gain at the three different frequencies corresponding to the first, centre and last channel of WiFi. From this table, it is also observed that the gain of the unit element has a higher variation in gain over the WiFi spectrum of 1.91 dB from channel 1 to channel 13. On the other hand, the gain variation of the array is more consistent at 0.43 dB, with a higher gain at the channel 7 of -4.71 dB.

Fig. 4-9 Gain comparison between unit element (dashed line) and array centre fed (solid line)

Table 4-2 Comparison of gain

	@2.412 GHz (dB)	@2.442 GHz (dB)	@2.472 GHz (dB)	Change in gain (dB)	Area (mm ²)
Unit element	-15.45	-16.14	-17.36	1.91	134
Array	-4.18	-4.28	-5.62	0.43	1600
Change	11.27	11.86	11.74	-	11.9X

Fig. 4-10(a) shows the simulated radiation pattern of the array in the x-z plane. The green rectangle represents the array, with the position and direction of the arrow indicating the top of the array and front of the array respectively. From this figure, it is also observed that the radiation pattern is not symmetrical. Further investigations in this section will analyse the effects on the radiation pattern, efficiency, gain, and reflection coefficient of feeding the array at different feed points.

Fig. 4-10 Simulated radiation patterns of array fed at the centre element of the array (a) x-z plane, and perspective view (b) channel 1 (c) channel 7 (d) channel13

4.3.2 Energy Harvesting Verification

Fig. 4-11 Energy harvesting test (a) using off the shelf router (b) using a signal generator

To test energy harvesting ability of the proposed array, two tests are conducted. Fig. 4-11(a) shows an experimental set up using an off the shelf router (TP-link TL-WR702N). The antenna array is placed at a distance of 30 cm. The measured resistance of the antenna is 250Ω . To achieve maximum power transfer, a load of 250Ω is soldered to the antenna array. The router is configured to transmit at a maximum power of 20 dBm.

According to the testing result, the received voltage is 30 mV with a corresponding power of $3.6 \mu\text{W}$. The efficiency of this set up calculated to be 0.0003%.

In another test (see Fig. 4-11(b)), a signal generator, which is connected to an amplifier and a monopole antenna is used. This experiment was conducted using the signal generator, connected to a RF amplifier (ADL5611-EVALZ from Analog Device). According to the specifications, the gain of the RF amplifier at 2.45 GHz is about 20 dB. This set up is connected to a monopole antenna placed 125 mm away from the proposed antenna array. The same load

of 250Ω is used. According to the testing result, the received voltage is 250 mV with a corresponding power of 250 μ W. The efficiency of this set up is about 10%.

The test set up is taken at distance of 30 cm between the test antenna and array. To establish an understanding of the efficiency of this antenna in the practical environment, we calculated the efficiency as the ratio of total power received compared to the total configured output power of the router or signal generator with RF amplifier.

In the two tests, the calculations do not include path loss and several assumptions are taken to simplify calculations. This gives us an indication on the performance of this 7 cell antenna array.

4.3.3 Analysis

From the last column in Table 4-1, a maximum area increase of 11.9 times from unit element to array is shown. According to (2.3), there should also be an increase in gain of the same magnitude when other parameters stay constant. However, from the observations in Table 4-2, the gain of the metaresonator array is 11.27 dB, 11.86 dB, and 11.74 dB higher than the unit element at the first, centre and last channel of WiFi respectively. This increase represents a 13.40 to 15.3 times improvement in gain.

From (2.3), it is known that other than directivity, the radiation efficiency also affects the gain. According to [52], we know that the conduction-dielectric efficiency or radiation efficiency, ϵ_{cd} can be expressed as

$$\epsilon_{cd} = \left[\frac{R_r}{R_L + R_r} \right] \quad (4.1)$$

where R_L is the conduction-dielectric losses, and R_r is the radiation resistance.

For a given metal conductor with a certain cross section, at frequencies above DC, according to [52], the high frequency resistance R_{hf} can be written as

$$R_{hf} = \frac{l}{P} \sqrt{\frac{\omega \mu_0}{2\sigma}} \quad (4.2)$$

where l is the length of the conductor. P is the perimeter of the cross section, ω is the angular frequency, μ_0 is the permeability of free-space, and σ is the conductivity of the metal.

For conductors of different shapes, R_L can be a function of R_{hf} . In the case of scaling a single ESA element to an array of seven element, R_L increases as the length of conductor increases. From (4.1), for ϵ_{cd} to increase, where R_L is significant for a small antenna, R_r has to increase by a larger amount than R_L .

From simulation results (see Table 4-3), it can be observed that the difference between the radiation efficiency and total efficiency is less than 0.5 dB. This difference which is known as the impedance mismatch loss, is an indication of how well the antenna is matched to the port. As both the unit element metaresonator antenna and the array have S_{11} of -25 dB and -26 dB respectively, the small impedance mismatch loss is expected.

Comparing the unit element with the array, it is observed that the radiation efficiency increases by between 2.78 dB to 3.55 dB. This shows that as the antenna size scales up, R_r increases by a larger amount than R_L , consistent with (4.1),

Table 4-3 Efficiency comparison between unit element and array

	2.412 GHz		2.442 GHz		2.472 GHz	
	ϵ_T (dB)	ϵ_{rad} (dB)	ϵ_T (dB)	ϵ_{rad} (dB)	ϵ_T (dB)	ϵ_{rad} (dB)
Unit Element	-8.76	-8.21	-9.22	-9.13	-10.99	-10.36
Array Centre Feed	-5.98	-5.62	-6.09	-5.73	-7.44	-7.08
Improvement	2.78	2.59	3.13	3.4	3.55	3.28

Next, investigating the radiation pattern, it is observed that due to the design of how each element is fed, and the interconnect of the elements of the array, the radiation pattern obtained when the array is fed in the middle element is not symmetrical (see Fig. 4-10). To understand how feeding of the array affects the behaviour of the array, the array is simulated at each of the different feed points. The S_{11} , gain, efficiencies and radiation patterns are compared. After each feed point has been positioned, the length of stubs is swept to determine the length of the stubs that will achieve resonance at the centre frequency. Table 4-4 lists the length of the stubs and S_{11} values with respect to the feed points.

From the table, it can be observed that the array is not as well matched at different feed points, but most of them at least met the minimum requirement of approximately -16 dB. The reason is that only the length of the stubs is modified, while the element separation distance remains at 80 mils. The reason for keeping the element separation constant this is that in real applications, after the arrays have been fabricated, the element separation cannot be changed, while the stub lengths can be trimmed to suit different feed points. Therefore, it is important to investigate the effects on the arrays at different feed points with the element separation distance kept constant.

Table 4-4 Location of feed point (bottom of PCB), length of stubs, and S_{11}

Feed point	Length of stubs (mm)	S_{11} (dB)
Centre	0.561	-28.2
Centre top	0.9	-15.8
Centre bottom	0.65	-16.3
Left top	0.76	-33.7
Left bottom	0.58	-19.3
Right top	0.75	-16.3
Right bottom	0.6	-19.2

Fig. 4-12 Gain comparison at different feed points

The gain is first investigated at different feed points (see Fig. 4-12). The gain plots when fed at the centre (in black), left (in blue), and right (in red). Those gain plots when fed at the top are in dashed lines, and bottom are in dotted lines. The initial feed point gain plot is in solid black line. From this figure, it is observed that when the array is fed at the sides (either left or right), the gain is higher than at the centre (either centre, top, or bottom). Among the sides, generally the gain from top fed is higher than the bottom fed.

From (2.3), the gain of an antenna is directly proportional to radiation efficiency ϵ_{cd} , and aperture efficiency ϵ_{ap} , and from [197], the definition of gain is the ratio of radiation intensity in a given direction to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically. Studying Table 4-5, it is observed that the gain obtained in Fig. 4-12 is directly related to the radiation efficiencies. This is consistent with (2.3) which explains that gain increases when the radiation efficiency of the antenna increases. The

differences in radiation efficiency can also be understood from (4.1) which states that the radiation efficiency is affected by changes in R_L and R_r . In this case, the radiation efficiency is changed when the array is fed at different points.

Next, Fig. 4-13 to Fig. 4-15 show the radiation patterns on the x-z plane for channels 1, 7 and 13 respectively. These plots are plotted separately to show the shape of the patterns otherwise not possible together due to the different scales. The first observation from the three figures is that the radiation pattern for each feed point is approximately similar for the three different frequencies. This shows that despite different extent of matching at different frequencies, the shape of the radiation pattern is consistent. Thus, only one perspective view of the radiation pattern is shown in Fig. 4-16 at channel 7. Also, feeding the array at the centre top element produces a pattern similar to feeding at the centre. However, when it is fed at the centre bottom element, the pattern is altered significantly to point to the 90° direction. This can be explained by the position of the feed point of the centre bottom element being closest to the bottom edge of the array with no other copper traces to affect the EM field generated.

Table 4-5 Simulated efficiencies

	2.412 GHz		2.442 GHz		2.472 GHz	
Feed point	ϵ_T (dB)	ϵ_{rad} (dB)	ϵ_T (dB)	ϵ_{rad} (dB)	ϵ_T (dB)	ϵ_{rad} (dB)
Centre	-5.98	-5.62	-6.09	-5.73	-7.44	-7.08
Centre Top	-10.25	-9.63	-10.10	-9.98	-10.98	-10.49
Centre Bottom	-6.92	-6.39	-7.64	-7.11	-9.19	-8.66
Left Top	-5.95	-5.58	-5.58	-5.22	-6.26	-5.89
Left Bottom	-6.08	-5.78	-5.58	-5.28	-6.29	-5.99
Right Top	-5.28	-4.67	-5.32	-4.71	-6.90	-6.29
Right Bottom	-6.12	-5.76	-5.50	-5.13	-6.15	-5.79

Fig. 4-13 Radiation pattern at channel 1 at different feed points (a) centre top (b) centre bottom (c) left top (d) left bottom (e) right top (f) right bottom

Fig. 4-14 Radiation pattern at channel 7 at different feed points (a) centre top (b) centre bottom (c) left top (d) left bottom (e) right top (f) right bottom

Fig. 4-15 Radiation pattern at channel 13 at different feed points (a) centre top (b) centre bottom (c) left top (d) left bottom (e) right top (f) right bottom

Fig. 4-16 Radiation pattern at channel 7 at different feed points (a) centre top (b) centre bottom (c) left top (d) left bottom (e) right top (f) right bottom

When the array is fed at the either side, it is also observed that the pattern is modified to point in different directions. When fed at the right or left top elements, the radiation pattern is pointed to the negative angles region, while when fed at the bottom, it is pointed to the positive angles region. During deployment in a building, this type of feeding allows additional flexibility in installation design such that the antenna arrays can be made to point in certain directions to improve capacity and spectral efficiency, while using the same antenna.

4.4 Concluding Remarks

To show the scalability of this electrically small metaresonator in forming an array using the concepts of MM, two different configurations are designed to demonstrate the flexibility of this approach. While both configurations are well matched at channel 7, having S_{11} of -20 dB and -26 dB for type A and type B configuration respectively, type B configuration, where the elements are in the same orientation with a scalable interconnection, is chosen for scalable applications in buildings. In addition, type B configuration has a larger -10 dB fractional bandwidth of 3.07% compared to type A of 0.33%. Using type B configuration, this array shows an improvement in gain of up to 11.86 dB and an increase in efficiency of up to 3.55 dB. Compared with other resonator-based arrays, where the element separations are 0.161λ [167] and 0.202λ [168], the proposed array combining the concepts of MM has the smallest element separation of 0.0165λ .

Owing to the stubbed design, the proposed array has an added advantage of tuning the resonant frequency to the required frequency at channel 7 by changing the stub length. Compared with the design in [168], where there is a deviation of 180 MHz from the initial planned 2 GHz due to some fixed dimension, this ability to tune improves the impedance matching of the array as it scales. Furthermore, as there are multiple feed points, the gain of the array can vary from -7.8 dB to -2.15 dB with changes in radiation pattern in terms of the direction of the main beam.

The contributions in this chapter include validating the concept of using SRR to design a 0.1λ

ESA to form an electrically small metaresonator, and combining concepts from MM and RF EH antenna array to create an array using this antenna as an element with the advantageous feature of frequency tuning. The objective of minimising element separation is also shown, with distances smaller than other designs. In addition, due to the specific manner which the elements can be fed and how they are interconnected, analysis on the gain, efficiencies and radiation patterns are done to demonstrate the flexibility of this design for use in the building context.

Chapter 5

Effects of Cement Paste Enhanced with Iron-Based Magnetic Particles on An Embedded Metaresonator Array

5.1 Overview

From chapter 3, we determined that it is possible to combine the concepts of SRR from MM with ESA and antenna array. In chapter 4, this concept was used to design an electrically small metaresonator. To demonstrate the scalability of this ESA as an element in an array, a

metaresonator array is designed and analysed. The results obtained show that performances of both the ESA and array are within expectations according to the design concepts of resonators. However, as the underlying principle of both the electrically small metaresonator and array is resonator based, the bandwidths are narrow at about 3%. As this put a limit on the operations of the antenna, it is desirable to increase its bandwidth. According to section 2.7, one way to increase the bandwidth is to modify the dielectric surrounding the antenna [176].

When the dielectric surrounding an antenna is changed, the performance of the antenna such as the matching and bandwidth also changes [176]. In the context of buildings, an antenna can be embedded in building materials for various applications such as SHM [180]–[182]. When an antenna is embedded into facades, walls, partitions, ceilings, and floors (mainly made up by cement and aggregates), the dielectric of cement will affect the properties of the embedded antenna. It is already known from chapter 2 that building materials such as concrete has high shielding effectiveness, and when antennas are embedded into cement paste, the radiated and received power are significantly reduced. However, there is little study to investigate how the type (metallic and/or magnetic particles) or size (micro or nano) can affect the antennas when they are embedded in terms of impedance matching and transmission properties of the materials.

In this chapter, we propose adding iron-based magnetic particles in cement paste to improve the EM transmission properties of concrete for embedded antennas. This particle is typically added to cement paste as an admixture to affect the colour of the concrete. To simulate embedded wireless applications, metaresonator arrays are embedded into two types of particles of two different sizes added to cement pastes. As a result of an improvement in the EM properties of the heterogeneous medium, we observed for all samples 1) an increase in the bandwidth of the embedded antenna compared to the antenna in air and 2) a shift in the resonance frequency that was smaller than that of an antenna in cement paste alone. In analysing the effects of different sizes of iron-based particles on the antennas, it was observed that those antennas embedded in cement paste enhanced with the addition of micro-sized particles, while having increased bandwidth, was able to maintain the general shape of the S_{11} curve compared to other samples and the control. In addition, for the required WiFi spectrum,

cement paste samples enhanced with micro-sized iron (III) oxide particles improved the transmission coefficient by as much as 10 dB.

5.2 Methodology

To determine the effects of the type of iron-based oxide particles have on embedded metaresonator array, two types of particles typically used as pigment admixtures are chosen; magnetite and iron (III) oxide. As the size of the particles affects the properties of the enhanced cement paste, two different size ranges (micro- and nanometer sizes) are also evaluated. With these different combinations of types of particles, and the different sizes, a total of four types of particles are used.

Magnetite of micro and nano-sized, and iron (III) oxide of micro and nano-sized particles are selected to add into the cement paste. Due to the availability of the particles in the market for pigment admixtures, micro-sized particles are taken as particles with a maximum dimension of less than 5 μm . For particles of nano-sized, however, each of the iron-based samples have different dimensions. However, in general, nano-sized particles has a maximum dimension of less than 100 nm or 50 nm. These sizes are determined by the manufacturer, which are the common sizes used in the industry. Details of the particles used are listed in Table 5-1.

Table 5-1 Types of particles, size, percentage by weight used and label

Sample type	Size	Percentage by weight	Label
Micro-sized magnetite	$< 5 \mu\text{m}$	0.5%	MAG-M
Nanosized magnetite	$5 - 100 \text{ nm}$	0.5%	MAG-N
Micro-sized iron (III) oxide	$< 5 \mu\text{m}$	0.5%	IOX-M
Nanosized iron (III) oxide	$< 50 \text{ nm}$	0.5%	IOX-N

To simulate the embedded antenna in cement paste, the proposed metaresonator antenna arrays designed in Chapter 3 are inserted in the centre of the samples. A total of fifty such sample arrays are prepared. Metaresonator arrays of similar S_{11} behaviours are selected for embedding in the cement pastes and comparison. Three cubic samples of 8 x 8 x 8 cm were produced for each mix with different particles, and cement paste alone as a control, as shown in Fig. 5-1(b). The samples are labelled in the following way. (type of particle)-(size of particle), where the type of particle can be magnetite (MAG), iron oxide (IOX), or control (CON); and the size of the particles can be micro- (M) or nanosized (N). Ordinary Portland cement was mixed with a water to cement ratio of 0.35 and the particles at 0.5 weight percentage. A superplasticizer was used at 260 ml/100 kg of cement to improve the workability of the cement mix. Fig. 5-1(a) shows the metaresonator array with coaxial RF cables soldered onto the feed points of the antenna. The cables extended out of the samples (see Fig. 5-1(c)) for measurements.

(a)

(b)

(c)

Fig. 5-1 Samples of antenna in enhanced cement paste (a) sample metaresonator array (b) preparation in moulds (c) all samples

These samples were cured for 3 days, before measurements are taken. While the samples were cured for 3 days, the effect of longer curing period is also studied, and the results are given in the Appendix A. Keysight N5242A PNA-X network analyser was used to measure the S_{11} of the samples via the coaxial cable, and a reference antenna was connected to another port to measure the S_{21} parameter. The experimental setups are shown in Fig. 5-2. S_{11} is measured with antenna embedded in cement paste samples connected to port 1 (see Fig. 5-2(a)), while S_{21} is measured with the reference antenna in air connected to port 1 and the antenna in cement paste samples connected to port 2 (see Fig. 5-2(b)).

The data plots of the S_{11} , captured within the frequency range of 2 to 3 GHz are displayed where the resonance frequency can be seen, together with the bandwidth and S_{11} values. The S_{21} measurement compares the amount of RF power that can be transferred from the reference antenna at port 1 to the embedded antenna connected to port 2. This result gives an indication on how the material surrounding the antenna affects the amount of RF power that can be transmitted into the material. As the metaresonator arrays were simulated and optimized in the CST® software in air, the S_{11} and S_{21} characteristics of the embedded metaresonators were shifted away from the resonant frequency of channel 7 (2.442 GHz) when embedded into the cement paste and enhanced cement pastes. This provides a good indication of the effects of the enhanced cement paste on an antenna with regards to frequency shift. Therefore, using these two plots, a comparison to assess the effects of small resonator antennas when embedded in building materials with and without the use of particles can be performed.

Fig. 5-2 Experimental setup (a) measurement of S_{11} , (b) measurement of S_{21}

5.3 Results

5.3.1 Effects on reflection coefficient

Fig. 5-3 shows the plots of measured S_{11} of the array in cement paste only (solid black line) and in other cement pastes enhanced with particles. The S_{11} of one sample antenna measured in air medium (solid gray line) is also shown together with the other plots to serve as a reference due to slight variations in the behaviour of each fabricated antenna. The frequency range between the two vertical dashed lines indicates the intended WiFi spectrum between channel 1 of 2.412 GHz to channel 13 of 2.472 GHz.

When the antenna was embedded in cement paste only (the solid black line), the behaviour of the antenna was significantly changed. At the designed centre frequency of 2.442 GHz, the S_{11} decreased by 13 dB. Within the 2 GHz to 3 GHz bandwidth, there was no noticeable resonance, although in general, the S_{11} value is lower than -10 dB. This result demonstrates that the impedance of the antenna was not well-matched throughout this bandwidth.

Fig. 5-3 Comparison of S_{11} of antenna array in air, and other cement paste enhanced particles from 2 to 3 GHz

To improve the performance of the embedded antenna in building materials, four types of iron-based magnetic particles were added to the cement paste mix: 1) micro-sized magnetite (solid red line), 2) nanosized magnetite (dashed red line), 3) micro-sized iron (III) oxide (solid blue line), and 4) nanosized iron (III) oxide (dashed blue line). Table 5-2 summarizes the effects that the different particles had on the embedded antenna. The change in S_{11} (in dB) and frequency shifts (in percentage, where a positive value indicates a shift to higher frequencies, and a negative value indicates a shift to lower frequencies) are compared to the measurement of the antenna in air. The change in S_{11} is taken as a decrease of the lowest S_{11} value within the

2 to 3 GHz frequency range of the antenna measured before and after embedding into the cement paste.

Table 5-2 Effects of media on S_{11} and resonance frequency

Sample type	Change in S_{11} (dB)	Frequency shift (%)
Cement paste only (Control)	-2.01	23.20
Cement paste with micro-sized magnetite particles	-17.26	-3.87
Cement paste with nanosized magnetite particles	-6.19	3.66
Cement paste with micro-sized iron oxide particles	-19.28	12.20
Cement paste with nanosized iron oxide particles	-9.07	8.35

For the control (cement paste only sample), the S_{11} value was decreased by 2.01 dB compared to the measurement done for the antenna in air. As there was no distinct dip in the S_{11} within the 2 to 3 GHz range, the lowest point of the S_{11} at 3 GHz was taken as the resonance frequency for comparison purposes. For cement paste enhanced with particles of nanometre size, it was observed that both samples enhanced with magnetite and iron (III) oxide cause the S_{11} values to decrease by 6.19 dB and 9.07 dB, respectively. For cement paste enhanced with particles of micrometre size, the magnetite and iron (III) oxide containing samples caused the S_{11} values to decrease by 17.26 dB and 19.28 dB, respectively. Comparing the samples enhanced with different types of particles, the cement paste samples enhanced with iron (III) oxide particles showed a decrease in the S_{11} larger than samples enhanced with magnetite particles. This indicates that the embedded antenna is better matched when surrounded by cement paste enhanced with iron (III) oxide particles. Comparing the samples enhanced with different particle sizes, the addition of micro-sized particles was able to decrease the S_{11} of the embedded antenna to values even better than the sample reference antenna measured in air. This finding shows that when the antenna was embedded into a cement paste enhanced with micro-sized magnetic iron-based particles, the S_{11} value was significantly better than the antenna in air.

Conversely, although cement pastes enhanced with nanosized magnetic iron-based could lower the S_{11} of the antennas, they were not able to improve this performance significantly to achieve results that were better than just the antenna in air. In all samples, the S_{11} was reduced to less than -10 dB within the intended WiFi bandwidth. A S_{11} value of -10 dB or less indicates a sufficient performance of the antenna where at least 90% of the input power is delivered to the antenna, and 10% is reflected.

Observing the shift in resonance frequency, assuming the resonance frequency of the control sample is at 3 GHz (lowest value within the test spectrum), the shift in frequency is 23.2%. For the cement paste sample enhanced with magnetite particles, the shifts in frequency is -3.87% and 3.66% for micro-sized and nano-sized particles respectively. Compared with the cement paste enhanced with iron (III) oxide particles samples, the shift is 12.20% and 8.35% for the micro-sized and nanosized particles respectively. In general, other than the sample of cement paste enhanced with micro-sized magnetite particles, the other three cement paste samples with particles had their resonance frequencies shifted to higher frequencies. In terms of the magnitude of shifting, the samples enhanced with iron (III) oxide particles had their resonance frequencies shifted more than those samples enhanced with magnetite particles. The cement paste sample enhanced with micro-sized iron (III) oxide particles shifted towards higher frequency by the most. A small shift in frequency shows that the surrounding material had slight detuning effect on the antenna.

When observing the overall shape of the S_{11} plots, it can be noted that the cement paste samples enhanced with micro-sized particles preserved the general shape of the S_{11} curves with a single distinct dip at the resonance frequency. The cement pastes enhanced with nanosized particles exhibited two resonance frequencies, altering the single dip at the resonance frequency shape of the antenna. In terms of bandwidth, both cement pastes enhanced with micro-sized particles increased the bandwidth of the antenna. The sample enhanced with magnetite particles had a larger effect than the sample enhanced with iron (III) oxide particles. For the nanosized particles, the double dip shape also had the effect of increased bandwidth when compared to the bandwidth of the reference antenna in air. From the observations of the S_{11} of the different

samples, enhanced cement paste with iron-based magnetic particles had a positive effect on the embedded antenna. The cement paste samples with micro-sized particles had a better effect than the cement pastes with nanosized particles on the antennas in terms of the decrease in S_{11} , bandwidth, and the shape of the S_{11} curve.

5.3.2 Effects on transmission coefficient

Fig. 5-4 shows the measured transmission coefficients of the samples taken between 2 and 3 GHz. The frequencies between the two vertical dotted lines indicates the intended WiFi spectrum. A reference measurement of the sample antenna used (the solid gray line) is also shown to indicate the performance of the antenna in air when not embedded.

Fig. 5-4 Comparison of S_{21} of antenna array in air, and other cement paste enhanced with particles from 2 to 3 GHz

At the intended frequency range of WiFi, the antenna measured in air has S_{21} values ranging from -30 dB to -37 dB. When the antenna was placed in the control (cement paste only) sample (the solid black line), the S_{21} was reduced by about 16 to 20 dB. This shows that the amount of RF power that can transmit through into the cement paste was significantly reduced. Focusing on the WiFi spectrum (within the 2 vertical dashed lines), when the antenna was placed in a cement paste enhanced with iron-based magnetic particles, most of the S_{21} curves improved except from 2.41 GHz to 2.46 GHz for the cement paste sample enhanced with micro-sized magnetite particles. Both cement pastes enhanced with nanosized particles had the same behaviour within this range.

For cement pastes enhanced with iron (III) oxide particles, the improvement of the S_{21} curve was the most significant. From Table 5-3, which shows a summary of the S_{21} within the WiFi spectrum, the cement paste enhanced with micro-sized magnetite produced the least improvement, with values at certain frequencies performing worse than the control sample (cement paste only). On the other hand, the cement pastes enhanced with iron (III) oxide sample presented the most improvement (up to 10.33 dB), observed at Channel 13 (2.472 GHz) of the WiFi band. This finding shows that by enhancing the cement paste with iron-based magnetic particles, the amount of RF power passing through the material increases compared to cement paste alone.

For frequencies lower than Channel 1 of WiFi (the left-hand region of the left vertical dashed line), those samples enhanced with iron (III) oxide particles showed a better S_{21} values when compared to the control, and samples enhanced with magnetite particles. At frequencies higher than Channel 13 of WiFi (the right-hand region of the right vertical dashed line), those samples enhanced with nanosized magnetite particles showed a poorer effect on S_{21} compared to the other samples.

Table 5-3 Effects of media on S_{21} within WiFi spectrum

Sample type	S_{21} (dB)		Change in S_{21} (dB)	
	Channel 1	Channel 13	Channel 1	Channel 13
Cement paste only (Control)	-46.47	-54.21	-	-
Cement paste with micro-sized magnetite particles	-53.23	-49.46	-6.76	4.75
Cement paste with nanosized magnetite particles	-45.10	-49.35	1.37	4.75
Cement paste with micro-sized iron oxide particles	-40.49	-43.88	5.98	10.33
Cement paste with nanosized iron oxide particles	-45.20	-49.52	1.27	4.69

5.4 Discussion

5.4.1 On the effects of quality factor and bandwidth

When the medium surrounding an antenna is changed, its behaviour also changes due to the different characteristic impedances. The material of the antenna develops capacitance and/or inductance which are coupled to its surrounding medium. Any metallic, magnetic, and/or dielectric material in the near field affects these inductive and capacitive values. When discussing about small resonant antennas, such as metaresonators, the quality factor is important, as they typically have small bandwidths, BW . The quality factor as provided by the Chu-limit [198] is written as

$$Q > \frac{1}{(ka)^3} + \frac{1}{ka} \quad (5.1)$$

where $k = \frac{2\pi}{\lambda}$, and a is the radius of the sphere enclosing the antenna.

In general, the quality factor [80] can be written as

$$Q = \frac{\omega W}{P_{rad}} \quad (5.2)$$

where ω is the angular frequency, W is the average stored energy, and P_{rad} is the radiated power.

It is beneficial that Q is small as BW is inversely proportional to Q , by having as little energy stored in the antenna, and thereby increasing the power radiated. One way to decrease the Q and improve the BW is to analyse the antenna using series and parallel resonance transmission lines (TLs). If a transverse electromagnetic is considered, TEM wave propagation on a two-wire conductor, and assuming sinusoidal steady-state, the complex propagation constant can be obtained as

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (5.3)$$

where α is the attenuation constant, $\beta = k_0\sqrt{\mu_r\epsilon_r}$ is the phase constant, $k_0 = \sqrt{\mu_0\epsilon_0}$ is the propagation constant (wave number) of a plane wave in free space, R is the series resistance per unit length ($\frac{\Omega}{m}$), L is the series inductance per unit length ($\frac{H}{m}$), G is the shunt conductance per unit length ($\frac{S}{m}$), and C is the shunt capacitance per unit length ($\frac{F}{m}$). From this perspective, the quality factor can be written as

$$Q = \frac{\beta}{2\alpha} = \frac{k_0\sqrt{\mu_r\epsilon_r}}{2\alpha} \quad (5.4)$$

Therefore, to lower the Q , loading the antenna with materials that have lower relative permittivity, ϵ_r , and relative permeability, μ_r is possible. From the findings, the Q is decreased as the BW increased for all four samples. The cement paste enhanced with micro-sized iron-

based magnetic particles improved the bandwidth of the antenna by a greater extent while maintaining the shape of the S_{11} curve. Comparing the two cement pastes enhanced with micro-sized particles samples, iron (III) oxide had a better effect on S_{21} for frequencies between 2 to 2.54 GHz, which suggests that this sample allows more RF power to transmit through into the material. Within the intended WiFi spectrum, the cement paste enhanced with micro-sized iron (III) oxide particles also allows the most RF power to transmit into the sample.

5.4.2 On the effects of impedance

Matter is conventionally considered to be made up of charged particles such as protons and electrons, and uncharged particles like neutrons. EM field can also be viewed as being made up of photons. When the field is applied to matter, the charged particles can respond in several ways. For example, a single charged particle will experience Lorentz force, F in the presence of an EM field given by

$$F = qE + qv \times B \quad (5.5)$$

where q is the charge of the particle, E is the electric field, B is the magnetic field and v is the velocity of the particle. In other ways, a material made up of many neutral particles may split up to form induced dipole moments that can align to the applied field.

In addition, the shape of the particles also has a significant impact on the effective permittivity of the material. The observation in [212] states that the mutual interaction between the typical circular cylinders and spheres is very different from that of square cylinders and cubes. The latter interaction caused a decrease in their dipole moments, and hence the effective permittivity.

Therefore, although the wavelength of the applied EM field used here is much larger than the size of both sizes of particles, the effects produced are very much different. This is mainly due to the additional interactions of the nanosized particles with the surrounding medium and embedded antenna which then contributed significantly to the effects on the impedance of embedded antenna.

It is commonly known that relative permittivity of a material is dependent on frequency [211]

$$\varepsilon_r(\omega) = \varepsilon'_r(\omega) - i\varepsilon''_r(\omega) \quad (5.6)$$

In addition, from [213], it is also shown that that complex permittivity and complex permeability varies significantly depending on the frequency, especially when the concentration of iron oxide particles increases. Thus, the readings we obtained are limited to the measured frequency span.

From the simulated results in [214], when an antenna is embedded in concrete, the S_{11} changes relative to the background permittivity. As the background permittivity is reduced or increased, the S_{11} is increased or decreased respectively. Comparing to our results, it shows a consistency. When nanosized particles are added, the S_{11} are reduced for the two samples. The shape of the S_{11} curve is also similar to the one shown in [214] when the background permittivity is reduced. This indicates that the type of nanosized particles that are added are in the form of square cylinders and cubes, which the applied EM field interactions caused a decrease in their dipole moments, and hence the effective permittivity as stated in [212]. From the measured data in [212], we also see that by adding micro-sized particles, it increases the permittivity, and hence S_{11} of both samples are decreased.

5.5 Concluding Remarks

Electrically small resonator-based antennas have narrow bandwidth. This is a well-known disadvantage of such antennas. To increase the bandwidth, an understanding was established to increase the power radiated and reduce the energy stored in the antenna [80]. In addition, by expressing the quality factor in terms of attenuation and phase constants, the quality factor can be reduced by either decreasing the relative permittivity or relative permeability, or both. Using this understanding, this chapter explores the possibility of using common admixture pigment used by the industry for colouring concrete or mortar to increase the bandwidth. Specifically,

iron-based oxides are chosen with two different sizes; micro-sized and nano-sized due to their low cost and availability.

With respect to S_{11} , cement paste samples enhanced with micro-sized iron-based particles have a positive effect on the impedance of the embedded metaresonator arrays. On the other hand, cement paste samples enhanced with nano-sized iron-based particles negatively affects the S_{11} of the embedded antennas such that the performances are worse than that of the antenna in air. Among the two types of iron-based particles, both samples enhanced with micro-sized particles increase the bandwidth of the embedded antennas with the S_{11} below -10 dB in general. To make a comparison between the two, if -20 dB bandwidth is measured, the sample enhanced with magnetite increases the -20 dB bandwidth of the array by about 2.3 times more than the sample enhanced with iron (III) oxide. In terms of frequency shift, the sample enhanced with micro-sized magnetite also produces a smaller shift towards the lower frequency than the sample enhanced with micro-size iron (III) oxide, which shift towards the higher frequency.

Next, examining the S_{21} , within the WiFi spectrum and to a larger extent of frequency range of between 2.2 GHz to 2.65 GHz, the sample enhanced with micro-sized iron (III) oxide particles produces the highest the transmission coefficient of the embedded metaresonator array with values even higher than the cement paste only control sample. This significant finding shows that by enhancing cement paste with micro-sized iron (III) oxide particles, the transmission of RF power into the material can be increased.

Considering the combined effects of the iron-based particles on S_{11} and S_{21} of the embedded metaresonator arrays, it can be concluded from these findings that enhancing cement paste with micro-sized iron (III) oxide is able to 1) increase the bandwidth of embedded metaresonator array while maintaining the overall shape of the S_{11} , 2) produces a smaller shift in resonant frequency than the control sample, and 3) increases the RF power transmission into the material.

The contribution in this chapter established the possibility of using commonly used admixture pigments of iron-based particles to affect the performance of the embedded metaresonator array

and RF transmission property of the material. More specifically, micro-sized iron (III) oxide has been identified to produce a balanced positive effect on both the embedded metaresonator array and building material.

Chapter 6

Effects of Concentration of Iron (III) Oxide Enhanced Cement Paste on Embedded Metaresonator Array in Building Materials

6.1 Overview

From the previous chapter, cement paste enhanced with micro-sized iron (III) oxide has been identified to have the best balance between the effects of increasing the bandwidth and

improving the transmission coefficient within the intended frequency over control sample. The weight percentage of particles mixed is 0.5% of the cement used.

Following these initial experiments, the next set of experiments is to determine how the concentration of micro-sized iron (III) oxide particles when added to cement paste affect the embedded metaresonator. Three samples of each concentration are produced, namely; 0% (control), 1%, 2%, 3%, and 4% iron (III) oxide mixed in cement paste.

Typically, in the industry, metal oxides is a type of admixture pigment added to cement paste to affect the colours of the cement [199][200]. The percentage by weight of pigment is regulated to be of less than ten percent according to American Standard of Testing and Materials, ASTM C 979-99. Increasing the amount of pigment will affect the strength of the concrete or mortar. In terms of colours, iron (III) oxide pigment is a type of pigment commonly used to change the colour of concrete or mortar from pink to red to maroon, depending on the concentration. Although the maximum percentage by weight of pigment used should not exceed ten percent, in practice, the amount of pigment used is between three to four percent. Any amount exceeding 4% will produce diminishing colour changes [201]. In this chapter, therefore, the amount of iron (III) oxide that are added to cement paste to form the samples are from 1% to 4% concentration by weight to reflect the typical amount of iron (III) oxide pigments used.

The finding in this chapter contributes to the knowledge of future embedded antennas design in recognising how the different typical concentrations of iron (III) oxide pigments when mixed to cement paste affects the performance of the embedded metaresonator array and resulting material.

6.2 Methodology

A total of 15 samples is fabricated with the labelling of the samples as shown in Table 6-1. Using similar method and set up as described in Chapter 5, these 15 samples are formed using the same moulds (see Fig. 5-1(b)) as before. Metaresonators array of the proposed type are

inserted in the middle of the samples (see Fig. 6-1(a)), and are cured for 7 days (see Fig. 6-1(b)). It is worth noting that as the concentration of iron (III) oxide increases, the shade of red also increases as expected of the pigment. S_{11} and S_{21} measurement methods are also the same with the previous chapter. The labelling of sample # and Ant## are used interchangeably in this chapter.

Table 6-1 List of samples fabricated

	0% (Control)	1%	2%	3%	4%
Sample No.	1	7	4	10	13
	2	8	5	11	14
	3	9	6	12	15

(a)

(b)

Fig. 6-1 Samples with metaresonator arrays inserted (a) Preparation (b) Cured

6.3 Results

6.3.1 Measurement results

The individual S_{11} and S_{21} measurement results of all the samples are shown in Appendix B. To compare the effects between the samples, metaresonators with similar S_{11} are selected. Table 6-2 summarizes the S_{11} at the resonating frequency, and maximum S_{21} value.

Table 6-2 Summary of S_{11} and S_{21} of samples

Sample Type	Sample No.	S_{11} (dB)		S_{21} (dB)	
		Before	After	Before	After
Control	Ant01	-23.70	-28.35	-10.48	-22.91
	Ant02	-20.28	-21.21	-12.22	-24.19
	Ant03	-26.95	-24.17	-15.31	-24.73
1% iron (III) oxide	Ant07	-20.82	-20.53	-10.69	-21.77
	Ant08	-23.02	-27.53	-16.97	-21.74
	Ant09	-21.05	-25.21	-12.36	-24.60
2% iron (III) oxide	Ant04	-33.49	-27.97	-14.55	-20.77
	Ant05	-21.60	-25.03	-11.56	-23.23
	Ant06	-15.72	-26.63	-14.06	-19.54
3% iron (III) oxide	Ant10	-30.89	-24.24	-11.72	-23.38
	Ant11	-26.69	-25.52	-13.70	-23.95
	Ant12	-18.65	-21.78	-14.70	-24.38
4% iron (III) oxide	Ant13	-27.89	-43.90	-13.60	-19.73
	Ant14	-33.51	-17.61	-13.75	-19.27
	Ant15	-29.85	-38.96	-11.84	-28.86

6.3.2 Selection of samples for comparison

Due to the nature of embedding the antennas into the cement paste, the performances of the embedded arrays are assessed before and after embedding. The samples are compared among each other within same concentration category. An observation of the S_{11} of the samples is done to select the samples that may be possibly affected during the embedding. Fig. 6-2 to Fig. 6-6 show the combined S_{11} plots before and after embedding in each of the control, and four different concentrations of iron (III) oxide enhanced samples. Fig. 6-2 shows the comparison of Ant01, Ant02, and Ant03 samples, Fig. 6-3 shows the comparison of Ant04, Ant05, and Ant06 samples, Fig. 6-4 shows the comparison of Ant07, Ant08, and Ant09, Fig. 6-5 shows the comparison of Ant10, Ant11, and Ant12, while Fig. 6-6 shows the comparison of samples Ant13, Ant14, and Ant15.

From these observations, in the control (see Fig. 6-2(b)), 1% (see Fig. 6-3(b)), and 2% (see Fig. 6-4(b)) iron (III) oxide enhanced cement paste samples, they exhibit similar S_{11} plots after embedding. This shows that after the antennas were inserted, the embedded antennas were affected similarly. All these nine samples are used for analysis and discussions. However, when we studied the 3% (see Fig. 6-5(b)) and 4% (see Fig. 6-6(b)) iron (III) oxide enhanced cement paste samples, samples Ant10 and Ant14 do not exhibit similar S_{11} plots as the other two samples in their categories. This shows that some degree of damaged may have occurred during the insertion. As a result, Ant10 and Ant14 will not be used for comparison in this study.

From the rest of the samples, an observation into S_{11} plots is done to select the suitable samples to compare with the controls from Table 6-2. They are summarised in Table 6-3.

Table 6-3 Summary of selection of samples for comparison

	Control	1%	2%	3%	4%
Sample	2	9	5	-	-
	3	-	-	11	13

(a)

(b)

Fig. 6-2 S_{11} comparison between samples 1, 2 and 3 (a) before embedding (b) after embedding

(a)

(b)

Fig. 6-3 S_{11} comparison between samples 4, 5 and 6 (a) before embedding (b) after embedding

(a)

(b)

Fig. 6-4 S_{11} comparison between samples 7, 8 and 9 (a) before embedding (b) after embedding

(a)

(b)

Fig. 6-5 S_{11} comparison between samples 10, 11 and 12 (a) before embedding (b) after embedding

(a)

(b)

Fig. 6-6 S_{11} comparison between samples 13, 14 and 15 (a) before embedding (b) after embedding

6.3.3 Comparison of antennas before embedding

With the antennas Ant02 (control), Ant05 (2%), and Ant09 (1%) selected for one group of comparison, and Ant03 (control), Ant11 (3%), and Ant13 (4%) selected for another group of comparison, a study of the S_{11} and S_{21} of the antennas is first done before they are inserted into the cement pastes. The comparisons are separated into two groups as the S_{11} of Ant02, Ant05, and Ant09 are similar, while Ant03, Ant11, and Ant13 are similar.

Fig. 6-7(a) shows the S_{11} comparison of the three antennas (Ant02, Ant05, and Ant09) before they are embedded into the cement paste. The resonant frequency of these three samples are very close with S_{11} of less than -20 dB, and the measurements between channel 1 and channel 13 are below -10 dB. The comparison of these three antennas before embedding in terms of S_{21} is shown in Fig. 6-7(b). These three antennas generally exhibit similar trend comparable to each other. Overall, within the WiFi band, the S_{21} is about -13 dB. These values deviate from the value of about -30 dB in Fig. 5-4. This difference may be due to different environment and period of measurement. However, this difference will not affect the results in this chapter as the emphasis is on the comparison study within the samples prepared in this set of experiments.

Fig. 6-8(a) shows the S_{11} comparison of the three antennas (Ant03, Ant11, and Ant13) in another group before they are embedded into the cement paste. The resonant frequency of these three samples are very close with S_{11} of less than -25 dB, and the measurements between channel 1 and channel 13 are below -10 dB. The comparison of these three antennas before embedding in terms of S_{21} is shown in Fig. 6-8(b). The S_{21} behaviour of the three antennas are also quite close to each other. Overall, within the WiFi band, the S_{21} is about -15 dB.

From these comparisons, the S_{11} and S_{21} of the selected antennas in the two groups to be embedded into the cement paste of different concentrations are very similar. This gives a good indication of the similarities of the antennas before embedding, which in turn, gives a good representation of the resulting effects after embedding these metaresonator arrays into the cement pastes.

(a)

(b)

Fig. 6-7 Comparison between samples 2, 5 and 9 before embedding of (a) S_{11} (b) S_{21}

(a)

(b)

Fig. 6-8 Comparison between samples 3, 11 and 13 before embedding of (a) S_{11} (b) S_{21}

6.4 Discussions

6.4.1 Effects on reflection coefficient

Fig. 6-9 shows the comparison of S_{11} among the 6 samples. From Fig. 6-9(a), when compared to the control sample (Ant02), adding 1% and 2% iron (III) oxide reduces the S_{11} by about 5.8 dB at 2.45 GHz. This behaviour is approximately the same for both samples with S_{11} values of -14.4 dB. In both samples, the S_{11} is also lower than the control from 2 GHz to about 2.8 GHz. At higher frequencies, the S_{11} for the control sample is lower than the other 2 samples.

When observing the comparison of S_{11} between the control (Ant03), and 3% and 4% iron (III) oxide samples, Ant11 sample displays a similar behaviour as samples Ant05 and Ant09. A very different behaviour is observed for the 4% iron (III) oxide sample. This sample produces a lower S_{11} by as much as 20 dB when compared with cement paste only control sample. This shows that cement paste enhanced with 4% iron (III) oxide improves the S_{11} property of the embedded antenna by the most amount, while maintaining the overall shape of the S_{11} curve.

Table 6-4 summarizes the change in the minimum S_{11} and the shift in frequency of the four samples compared to the control, while Fig. 6-10 shows the improvement of the iron (III) oxide enhanced samples over the control. While all four iron (III) oxide enhanced samples show improvement over the control sample, 1% to 3% iron (III) oxide concentration samples show an improvement significantly lesser than 4% iron (III) oxide concentration sample. In terms of the shift in resonant frequency, the 4% iron (III) oxide concentration sample is also the only sample that shifts the resonant frequency to a higher frequency. From these results, 4% iron (III) oxide concentration sample improves the S_{11} by the most (20 dB) with a small shift of 70 MHz compared to the other samples. A combined graph of samples Ant03, Ant13, and Ant15 (see Fig. 6-11) shows that antennas with similar S_{11} before embedding behaved similarly after embedding. This further confirms that 4% iron (III) oxide concentration improves the S_{11} of the embedded antenna the most and shifts the resonant frequency by approximately 70 MHz.

(a)

(b)

Fig. 6-9 Comparison of S_{11} after embedding between (a) samples 2, 5 and 9 (b) samples 3, 11 and 13

Table 6-4 Effects of concentration of iron (III) oxide particles on S_{11} and resonance frequency

Sample type	Change in S_{11} (dB)	Frequency shift (MHz)
Cement paste with 1% micro-sized iron oxide particles	-4	-90
Cement paste with 2% micro-sized iron oxide particles	-3.82	-110
Cement paste with 3% micro-sized iron oxide particles	-1.36	-40
Cement paste with 4% micro-sized iron oxide particles	-19.73	70

Fig. 6-10 Improvement in S_{11} of iron (III) oxide enhanced samples over control

In terms of comparing the frequency shift from channel 7, the samples enhanced with 4% iron (III) oxide also shifted by the most, even more than the control. From this observation, it is also noted that by adding less than 4% iron (III) oxide particles, the frequency shift effect is similar to adding nano-sized iron-based particles to cement paste.

(a)

(b)

Fig. 6-11 Comparison of S_{11} between samples Ant03, Ant13, and Ant15 (a) before embedding
(b) after embedding

6.4.2 Effects on transmission coefficient

Fig. 6-12 shows the measured S_{21} of the samples taken between 2 to 3 GHz. In Fig. 6-12(a), the S_{21} plot of the control (Ant02) is compared with the other two samples of 1% iron (III) oxide concentration (Ant09) and 2% iron (III) oxide concentration (Ant05). The Ant02 sample and Ant09 sample closely resembles each other within the WiFi spectrum from 2.442 to 2.472 GHz. In addition, from 2 GHz to 2.74 GHz, the plots of S_{21} of Ant02 and Ant09 samples are also similar. From 2.74 GHz to 2.88 GHz, however, Ant09 shows a better S_{21} than the control. From 2.88 GHz to 3 GHz, the control has the highest S_{21} values. Comparing with Ant05, the S_{21} plot from 2 GHz to 2.4 GHz is similar to the other two samples. From 2.4 GHz to 2.7 GHz, the S_{21} of Ant05 shows higher S_{21} values than the control and 1% iron (III) oxide concentration samples. For frequencies higher than 2.7 GHz, the trend is similar to Ant09 sample, while having lower S_{21} values from frequencies 2.84 GHz and higher. From this graph comparison, 2% iron (III) oxide concentration sample has a better S_{21} behaviour from 2 GHz to 2.7 GHz, which covers the WiFi spectrum, than the control and 1% iron (III) oxide concentration sample.

To compare other iron (III) oxide concentration samples, the S_{21} plots of another control (Ant03) with 3% and 4% iron (III) oxide concentration samples are shown in Fig. 6-12(b). Comparing the control with 3% iron (III) oxide concentration sample, the S_{21} plot is similar for both within the WiFi frequency, and between 2.4 GHz to 2.53 GHz. At frequencies below 2.4 GHz, Ant11 and Ant13 samples have relatively similar trends. These trends are also similar to the control, with the exception that the S_{21} of control sample is shifted to the lower frequency end by about 60 MHz. Within the frequencies between 2.4 GHz to 2.53 GHz, Ant04 sample of 4% iron (III) oxide concentration has the highest S_{21} values compared to the other two samples. This also indicates that within the WiFi spectrum, cement paste enhanced with 4% iron (III) oxide particles has a more significant improvement on S_{21} compared to the control and 3% iron (III) oxide concentration samples. Between 2.6 GHz to 2.8 GHz, the S_{21} trend of the Ant03 and Ant11 are similar, although the control sample has a higher S_{21} value range. Within this frequency, Ant13 sample has a mixture of higher and lower S_{21} values than the control and

Ant11 at different frequency spans. From 2.8 GHz to 3 GHz, the S_{21} plots of Ant11 and Ant13 are similar in both magnitude and trend. Compared with the control, both Ant11 and Ant13 have a worse S_{21} range. Within this range of frequencies, the control has a better impact on the S_{21} of the embedded antenna than the 3% and 4% iron (III) oxide concentration samples.

From Fig. 6-12, it is observed that at the WiFi centre frequency of 2.45 GHz, the S_{21} values of the two control samples Ant02 and Ant03 are similar at about -25 dB. With this value, the S_{21} values of all the samples at this centre frequency can be compared. To compare at this frequency, the effect of the concentration of iron (III) oxide has on the samples, a plot of the differences in the S_{21} values of the samples compared with their respective controls are shown in Fig. 6-13. A positive difference indicates a better S_{21} value over the controls, while a negative difference shows a worse S_{21} value. To begin comparison, the S_{21} of the arrays before embedding are compared to gain an understanding of the improvement of when they are embedded in different types of cement pastes. From Fig. 6-13(a), it is clear that at the centre frequency, the S_{21} of the different arrays, despite having similar S_{11} , have a difference of between -0.88 dB to 2.63 dB over the control arrays.

From Fig. 6-13(b), the findings show that all the samples with iron (III) oxide enhanced cement paste display an improvement over the controls except for 1% iron (III) oxide concentration sample. The magnitude of improvement is about 1 dB for samples with 2%, and 3% iron (III) oxide concentration. For the 4% iron (III) oxide concentration sample, however, the effect on the S_{21} is significantly higher at more than 5 dB. This result suggests that by adding 4% iron (III) oxide to cement paste, the effect on S_{21} of the embedded antenna is the highest among the other concentrations and control.

Additional interest is to analyse S_{21} values within the WiFi spectrum. Table 6-5 summarizes the S_{21} values of the measured values at channel 1 and channel 13 after embedding in the column labelled S_{21} (dB).

(a)

(b)

Fig. 6-12 Comparison of S_{21} after embedding between (a) samples 2, 5 and 9 (b) samples 3, 11 and 13

(a)

(b)

Fig. 6-13 S_{21} difference over control samples at 2.45 GHz (a) before embedding, (b) after embedding

In this table, the magnitude of change over the controls are also calculated to indicate the amount of change over the controls in the last two columns. Similarly, a positive difference indicates a better S_{21} value over the controls, while a negative difference shows a worse S_{21} value. The sample with 1% concentration of iron (III) oxide shows a slight decrease in S_{21} values, while the samples with 2% and 3% concentration of iron (III) oxide show an improvement of less than 1.2 dB. More significantly, cement paste enhanced with 4% iron (III) oxide shows the most improvement by more than 7 dB.

Table 6-5 Effects of concentration of iron (III) oxide on S_{21} within WiFi spectrum

Sample type		S_{21} (dB)		Change in S_{21} over control (dB)	
Description	Sample No.	Channel 1	Channel 13	Channel 1	Channel 13
Cement paste with 0% iron (III) oxide (Control)	Ant02	-27.38	-27.00	—	
	Ant03	-26.04	-30.04		
Cement paste with 1% iron (III) oxide	Ant09	-27.73	-27.58	-0.35	-0.57
Cement paste with 2% iron (III) oxide	Ant05	-26.47	-25.85	0.91	1.15
Cement paste with 3% iron (III) oxide	Ant11	-25.47	-28.98	0.57	1.06
Cement paste with 4% iron (III) oxide	Ant13	-22.73	-22.76	3.32	7.29

6.4.3 Effects of concentration of iron (III) oxide on S_{11} and S_{21}

From Fig. 6-10 and Fig. 6-13(b), it is noticed that the relationships between the concentration of iron (III) oxide and improvement of S_{11} and S_{21} are not linear. This behaviour can be explained by understanding how concentration of iron (III) oxide affects the total permittivity of the medium, and how permittivity affects S_{11} and S_{21} .

From Nicholson-Weir-Ross method [215], we know that S_{11} can be written as

$$S_{11} = \frac{\Gamma(1 - T^2)}{(1 - \Gamma^2 T^2)} \quad (6.1)$$

and S_{21} can be written as

$$S_{21} = \frac{T(1 - \Gamma^2)}{(1 - \Gamma^2 T^2)} \quad (6.2)$$

where the reflection coefficient is given as

$$\Gamma = X \pm \sqrt{X^2 - 1}, |\Gamma| < 1 \quad (6.3)$$

where

$$X = \frac{S_{11}^2 - S_{21}^2 + 1}{2S_{11}} \quad (6.4)$$

and the transmission coefficient can be written as

$$T = \frac{S_{11} - S_{21} - \Gamma}{1 - (S_{11} + S_{21})\Gamma} \quad (6.5)$$

The permeability can be given as

$$\mu_r = \frac{1 + \Gamma}{\Lambda(1 - \Gamma) \sqrt{\frac{1}{\lambda_0^2} - \frac{1}{\lambda_c^2}}} \quad (6.6)$$

where λ_0 is the free space wavelength, λ_c is the cut-off wavelength and

$$\frac{1}{\Lambda^2} = \left(\frac{\epsilon_r * \mu_r}{\lambda_0^2} - \frac{1}{\lambda_c^2} \right) = - \left(\frac{1}{2\pi L} \ln \left(\frac{1}{T} \right) \right)^2 \quad (6.7)$$

where L is the thickness of the sample.

The permittivity can be written as

$$\varepsilon_r = \frac{\lambda_0^2}{\mu_r} \left(\frac{1}{\lambda_c^2} + \frac{1}{\Lambda^2} \right) \quad (6.8)$$

From (6.8), we can obtain the transmission coefficient, T as

$$T = e^{\pi L \left(\frac{\varepsilon_r \mu_r}{\lambda_0^2} - \frac{1}{\lambda_c^2} \right)} \quad (6.9)$$

From (6.6), we can obtain the reflection coefficient, Γ as

$$\Gamma = \frac{\mu_r \Lambda \sqrt{\frac{1}{\lambda_0^2} - \frac{1}{\lambda_c^2}} - 1}{\mu_r \Lambda \sqrt{\frac{1}{\lambda_0^2} - \frac{1}{\lambda_c^2}} + 1} \quad (6.10)$$

The formulae shown in (6.1) and (6.2) for S_{11} and S_{21} show that they are dependent on T and Γ . From (6.9) and (6.10), both coefficients are dependant mainly on ε_r and μ_r , which are related to the material.

As the percentage of iron oxide increases from 0% to 4%, ε_r increases nonlinearly, with greater increase at higher percentages, and significantly more than μ_r [213]. Since S_{11} and S_{21} are dependent on ε_r and μ_r , and from (6.1) and (6.2), and (6.9) and (6.10), as the percentage of iron oxide increases, S_{11} and S_{21} changes, but more so at higher concentration of iron oxide.

6.5 Concluding Remarks

After identifying micro-sized iron (III) oxide particles to be most suitable admixture pigment to enhance cement paste for embedded metaresonator arrays, this chapter continues to study the effects of the percentage by weight of this admixture on the properties of the material and embedded antenna. A total of 15 metaresonator array samples are fabricated and arrays with similar S_{11} are grouped together. Using these arrays, a total of 15 cement paste samples are prepared and categorised into five groups of three samples. The categories of samples are cement paste only (control), and cement paste enhanced with 1% to 4% concentration of iron (III) oxide particles in increment of 1%. The array with the most similar performance in each group is selected and compared with the control in terms of S_{11} and S_{21} .

From the findings, in terms of S_{11} , all four concentrations of iron (III) oxide in cement paste improved the S_{11} by lowering the value of S_{11} by a magnitude of between -1.4 dB to -19.7 dB. This indicates that the arrays are better matched when iron (III) oxide particles are added when compared to just embedding the arrays in cement paste alone. With respect to the frequency shift, the addition of iron (III) oxide affects the resonant frequency by between 40 MHz to 110 MHz. The shift for the samples enhanced with 1% to 4% iron (III) oxide particles are similar to those of the nano-sized iron-based particles. The shift towards higher frequencies is most prominent with the samples enhanced with 4% micro-sized iron (III) oxide. Although the shift in frequency is the most, the improvement in magnitude of S_{11} is also the greatest by about 20 dB. The overall shape of the S_{11} is also maintained while the other samples are distorted when compared with the S_{11} plot of the array in air.

In terms of S_{21} , the findings show that adding 1% does not improve the S_{21} , while 2% and 3% micro-sized iron (III) oxide to cement paste slightly increase the amount of RF power that is transmitted within the WiFi spectrum. Significantly, cement paste enhanced with 4% micro-sized iron (III) oxide particles show a large improvement in the S_{21} within the WiFi spectrum of up to 7.29 dB over the control sample.

From the findings in this chapter, it suggests that for applications of embedded metaresonator arrays in building materials, when cement paste is enhanced with 4% micro-sized iron (III) oxide particles, the array can benefit by having 1) an improvement in the S_{11} , and 2) a preservation of the overall shape of the S_{11} . This provides a possibility of tuning the array to the desired frequency in the future. Another important observation from this finding also suggests that the RF transmission property of the resulting material is improved when cement paste is enhanced with 4% micro-sized iron (III) oxide.

The contribution in this chapter demonstrates that by mixing 4% of the commonly used admixture pigment of micro-sized iron (III) oxide in cement paste, the performance of the embedded metaresonator array and the RF transmission property of the building material within the WiFi spectrum are improved.

Chapter 7

Effects of 4% Concentration on Iron (III) Oxide Enhanced Plaster on Mortar on Embedded Metaresonator Array

7.1 Overview

From the previous two chapters, 4% concentration by weight of micro-sized iron (III) oxide has been identified to have the best effect on increasing the bandwidth, improving the S_{11} and improving the transmission coefficient within the intended frequency over the control sample.

In the context of buildings, plaster is commonly used on the surface of concrete to create a smooth finish. In this chapter, we investigate the effects of 1) embedding the metaresonator array in plaster, and 2) with the array placed on the surface of mortar covered with a layer of plaster. The knowledge gained here provides a background on the application of embedded antennas placed in façade and exterior wall finishing.

7.2 Methodology

A total of 8 samples are fabricated with the labelling of the samples as shown in Table 7-1. Two samples are prepared with plaster alone, while six samples are prepared with plaster on mortar. As there are no practical application where plaster is used alone in a building context, only two samples are prepared to develop a basic understanding on how plaster only and 4% iron (III) oxide enhanced plaster affect the embedded metaresonator array. Next, six plaster on mortar samples are prepared with three samples as controls having no iron (III) oxide particles added, while the other three samples are enhanced with 4% iron (III) oxide by weight. All the plaster used are mixed with cement and lime in the ratio of 1:1.

Using similar method and set up as described in Chapter 4, these eight samples are prepared using the same moulds (see Fig. 5-1(b)) as before. For the first two plaster only samples, the metaresonator array is inserted in the middle of the plaster paste (see Fig. 7-1(a)). The next six samples are prepared in a way different from the previous chapters.

Table 7-1 List of samples fabricated

	0% (Control)	4% iron (III) oxide	Remarks
Sample No.	23	25	Plaster only
	16	20	Plaster on mortar
	17	21	
	18	22

Fig. 7-1 Preparation of samples (a) preparation of plaster samples (b) placement of arrays onto wet cement paste for Ant20, Ant21, Ant22 (c) placement of arrays onto wet cement paste for Ant16, Ant17, Ant18 (d) drying of samples (e) plaster applied (f) dry samples

First, the six moulds are filled with cement paste to the top. Next, with the cement paste still wet, metaresonators with RF cables attached are laid on top of the cement paste as shown in Fig. 7-1(b) and Fig. 7-1(c). The samples are left to dry for three days (see Fig. 7-1(d)) before the six samples are layered with 1 cm of plaster or iron (III) oxide enhanced plaster (see Fig. 7-1(e)). The final samples (see Fig. 7-1(f)) are cured for another three days before data was collected. Fig. 7-1(e) and Fig. 7-1(f) also show that those samples without iron (III) oxide added are grey, and those with 4% iron (III) oxide added are coloured red.

7.3 Results

7.3.1 Measurement results – plaster only

To develop an understanding on how plaster affects the performance of embedded metaresonators as compared to cement paste done in the previous chapters, two samples of plaster only (Ant23 sample) and plaster enhanced with 4% iron (III) oxide (Ant25 sample) are measured. Fig. 7-2 shows the S_{11} of two metaresonator arrays before and after embedding into the cement paste. The two samples have similar minimum S_{11} value of -22 dB and -27 dB at frequency of 2.46 GHz and 2.44 GHz respectively. After embedding, the plaster only sample (Ant23) shifted the resonating frequency to 2.09 GHz while generally lowering the S_{11} values to below -10 dB. For the plaster with 4% iron (III) oxide sample, the resonating frequency is relatively unchanged at 2.45 GHz with a S_{11} value of -25.7 dB, down from -22.6 dB. This suggests that while the plaster only material shifted the resonating frequency to a lower frequency, the plaster enhanced with 4% iron (III) oxide generally maintained the overall shape of S_{11} and not affecting the resonating frequency while increasing the bandwidth.

In terms of S_{21} (see Fig. 7-3), before embedding (see Fig. 7-3(a)), the S_{21} are relatively similar from 2.14 GHz to 2.51 GHz. Lower frequencies show higher S_{21} values for Ant25, while generally higher S_{21} values from 2.54 GHz and higher frequencies. After embedding, from 2 GHz to 2.50 GHz, sample Ant23 generally has a higher S_{21} than Ant25. From 2.50 GHz to 3 GHz, sample Ant23 generally has higher S_{21} values than sample Ant25.

(a)

(b)

Fig. 7-2 S_{11} comparison between samples 23 and 25 (a) before embedding (b) after embedding

(a)

(b)

Fig. 7-3 S_{21} comparison between samples 23 and 25 (a) before embedding (b) after embedding

Despite the differences of about 7 dB across 2 GHz to 3 GHz range, the behaviour of S_{21} of the two samples follow similar trends. Another observation is that the maximum S_{21} value is decreased by about 11 dB after the metaresonator arrays are embedded into plaster only sample, while plaster enhanced with 4% iron (III) oxide samples has a 20 dB decrease.

Table 7-2 summarizes the minimum S_{11} and maximum S_{21} values with the 2 GHz to 3 GHz range before and after embedding the metaresonator arrays into the plaster materials.

Table 7-2 Summary of S_{11} and S_{21} of samples

Sample Type	Sample No.	Minimum S_{11} (dB)		Maximum S_{21} (dB)	
		Before	After	Before	After
Plaster only	Ant23	-26.99	-20.53	-13.35	-24.51
Plaster with 4% iron (III) oxide	Ant25	-22.55	-25.78	-11.63	-31.74

7.3.2 Measurement results – plaster on mortar

A total of six plaster on cement samples were prepared. Samples Ant16, Ant17, and Ant18 are samples with only plaster on mortar. Samples Ant20, Ant21, and Ant22 are samples with 4% iron (III) oxide enhanced plaster on mortar.

Fig. 7-4(a) shows the S_{11} of the three antennas before they are embedded for plaster only samples. Although there are some slight differences, the S_{11} values are generally in good agreement with each other. After embedding, all the three antennas are distorted significantly (see Fig. 7-4(b)). Within the WiFi range, the S_{11} is deteriorated to -6 dB or more beyond the range of acceptable S_{11} of an antenna. For frequencies higher than 2.6 GHz, all S_{11} of the three antennas decrease to below -6 dB and seem to resonate between 2.96 to 3 GHz, although sample Ant16 has S_{11} values generally below -6 dB for the 2 GHz to 3 GHz span.

(a)

(b)

Fig. 7-4 S_{11} comparison between samples 16, 17 and 18 (a) before embedding (b) after embedding

Fig. 7-5(a) shows the S_{11} of the three antennas before they are embedded for the 4% iron (III) oxide enhanced plaster samples. After embedding, as shown in Fig. 7-5(b), the behaviours of the S_{11} of all three samples are distorted significantly as those shown in Fig. 7-4(b). The two apparent differences are that the resonating frequencies are shifted to higher frequencies beyond 3 GHz, and the -6 dB S_{11} values for all three samples are 2.75 GHz and above, although sample Ant20 is generally less than -6 dB for the entire measurement range

In terms of S_{21} , Fig. 7-6 shows the S_{21} of samples Ant16, Ant17, and Ant18. Before embedding, the three antennas have similar S_{21} , although sample Ant16 has lower S_{21} values below 2.38 GHz (see Fig. 7-6(a)). After embedding, the S_{21} plots are very similar for frequencies less than 2.57 GHz. At frequencies higher than 2.57 GHz, the S_{21} plots of the three antennas start to deviate, but in general, still follow the same trend. Comparing Fig. 7-6(a) and Fig. 7-6(b), by embedding the metaresonator arrays into plaster only layer on mortar, it does not significantly change the S_{21} values within the WiFi spectrum. At frequencies lower than 2.4 GHz, the S_{21} values are generally lowered by 5 dB to 10 dB. At frequencies higher than 2.5 GHz, the S_{21} values are decreased by between 10 dB to 20 dB.

For samples Ant20, Ant21, and Ant22, before embedding (see Fig. 7-7(a)), the S_{21} of all the antennas generally follow a similar trend. Between frequencies of 2.4 GHz to 2.5 GHz, Ant21 produces higher S_{21} values than the other two samples with Ant20 having the lowest S_{21} values. The difference between the highest and lowest S_{21} values among these three samples in this frequency range is about 11 dB. Despite the difference within this frequency range, after embedding, all three antennas have much a similar trend and magnitude (see Fig. 7-7(b)). Different from the plaster only layer on mortar samples, the S_{21} is increased by between 0 dB (Ant21) to 7.7 dB (Ant20). For frequencies below 2.4 GHz and above 2.5 GHz, the effects of embedding the arrays in iron (III) oxide enhanced plaster layer on mortar produces similar effects as those shown in Fig. 7-6(b).

Table 7-3 summarizes the minimum S_{11} and maximum S_{21} values of the samples within the 2 GHz to 3 GHz range.

(a)

(b)

Fig. 7-5 S_{11} comparison between samples 20, 21 and 22 (a) before embedding (b) after embedding

(a)

(b)

Fig. 7-6 S_{21} comparison between samples 16, 17 and 18 (a) before embedding (b) after embedding

(a)

(b)

Fig. 7-7 S_{21} comparison between samples 20, 21 and 22 (a) before embedding (b) after embedding

Table 7-3 Summary of S_{11} and S_{21} of plaster on mortar samples

Sample Type	Sample No.	S_{11} (dB)		S_{21} (dB)	
		Before	After	Before	After
Plaster on mortar	Ant16	-31.38	-13.21	-13.59	-14.69
	Ant17	-23.50	-10.56	-11.00	-11.48
	Ant18	-25.96	-13.88	-11.71	-13.48
4% ion (III) oxide plaster on mortar	Ant20	-32.41	-13.68	-16.34	-13.26
	Ant21	-22.72	-10.43	-12.15	-14.24
	Ant22	-33.72	-15.27	-14.28	-12.52

7.3.3 Selection of samples for comparison – plaster only

In addition to comparing among the two plaster samples, two samples from the previous chapter are chosen for this comparison. The cement paste only (Ant03) and the 4% iron (III) oxide enhanced cement paste (Ant13) samples are selected with a S_{11} before embedding as close to the two plaster samples. Fig. 7-8(a) shows the S_{11} of the four samples before embedding. The S_{11} values of the metaresonator arrays are quite close for comparison. The S_{21} of these four samples are shown in Fig. 7-8(b). At the frequencies between 2.4 GHz to 2.5 GHz, the S_{21} of all four samples have very similar values. Outside of this frequency range, the S_{21} values of the four samples vary, but generally follow the same trend.

Table 7-4 summarizes the minimum S_{11} and maximum S_{21} values before and after embedding. These four samples will enable the development of an understanding on the differences between plaster only, cement paste only, plaster enhanced with 4% iron (III) oxide, and cement paste enhanced with 4% iron (III) oxide.

(a)

(b)

Fig. 7-8 Comparison between samples 23, 25, 03 and 13 before embedding (a) S_{11} , (b) S_{21}

Table 7-4 Summary of S_{11} and S_{21} of samples

Sample Type	Sample No.	S_{11} (dB)		S_{21} (dB)	
		Before	After	Before	After
Plaster only	Ant23	-26.99	-20.53	-13.35	-24.51
Cement paste only	Ant03	-26.95	-24.17	-15.31	-24.73
Plaster with 4% iron (III) oxide	Ant25	-22.55	-25.78	-11.63	-31.74
Cement paste with 4% iron (III) oxide	Ant 13	-27.89	-43.90	-13.60	-19.73

7.3.4 Selection of samples for comparison – plaster layer on mortar

From Table 7-3, samples with similar S_{11} are selected for comparison of the plaster on cement and 4% iron (III) oxide enhanced plaster on mortar. The four samples that are selected are listed in Table 7-5.

Table 7-5 Summary of selection of plaster on mortar samples for comparison

	Plaster on mortar (Control)	4% iron (III) oxide enhance plaster on mortar
Sample	Ant16	Ant20
	Ant17	Ant21

Fig. 7-9(a) shows the S_{11} plot of Ant16 and Ant20. These results show that the two metaresonator arrays are in close agreement before embedding. In terms of S_{21} , both antenna samples are also following similar trends despite having about 10 dB of difference at 2.45 GHz (see Fig. 7-9(b)). Fig. 7-10(a) also shows that Ant17 and Ant21 are very similar before embedding. The S_{21} plots (see Fig. 7-10(b)) are also having similar trend in general.

(a)

(b)

Fig. 7-9 Comparison between samples 16, and 20 before embedding (a) S_{11} (b) S_{21}

(a)

(b)

Fig. 7-10 Comparison between samples 17, and 21 before embedding (a) S_{11} (b) S_{21}

7.4 Discussions

7.4.1 Plaster only — effects on reflection coefficient

Fig. 7-11(a) shows the effects on S_{11} after embedding in the four different materials. The plaster only sample (Ant23) performed the worst, with the S_{11} graph distorted, although the values are generally below -10 dB. Comparing to cement paste only sample (Ant03), the shape of the S_{11} plot is still preserved, albeit shifted to a higher frequency of 2.77 GHz. Just by contrasting the two control samples, the plaster only sample (Ant23) shifted the resonating frequency to a lower region while the cement paste only sample (Ant03) shifted the resonating frequency to a higher one. Generally, the single dip shape of the S_{11} is preserved. The magnitudes of the shift are approximately similar at -37 MHz and 32 MHz for the plaster only sample and cement paste only sample respectively. The effect on S_{11} when the array is embedded in plaster only is poor while that in the cement paste only material is better. This suggests that inserting in plaster only material is not as suitable as cement paste only, although generally the values are less than -10 dB.

When 4% iron (III) oxide is added to plaster (Ant25) and cement paste (Ant13), the effects of this enhancement is more evident. For Ant25, the resonating frequency is not affected significantly while the S_{11} value at this point is decreased by 4.9 dB (see Fig. 7-2). This indicates that plaster enhanced with 4% iron (III) oxide improves the matching of the embedded array without altering the resonant frequency. Comparatively, the previous experimental data of Ant13 shows that cement paste enhanced with 4% iron (III) oxide shifted the resonating frequency to the higher one at 2.84 GHz. This suggests that 4% iron (III) oxide when added to plaster has much lesser impact on the matching of the embedded array than cement paste.

This effect can be understood by studying the composition of plaster which is made up by 50% cement and 50% lime. Lime is essentially calcium oxide, which has a static permittivity higher than that of iron oxide. However, when water is added to calcium oxide, the following reaction happens.

Upon setting when excess water has been evaporated, carbonation takes place

When the sample is set, the lime added to create plaster, turns into calcium carbonate. From [216], the dielectric constant of calcium carbonate is between 1.8 to 2.0, which is lower than cement of between 2.5 to 7.0 [206][207]. When materials of different types are mixed, the effective permittivity can be obtained by using Maxwell-Garnet mixture equation [211]

$$\frac{\varepsilon'_{eff} - \varepsilon'_2}{\varepsilon'_{eff} + 2\varepsilon'_2} = \theta \frac{\varepsilon'_1 - \varepsilon'_2}{\varepsilon'_1 + 2\varepsilon'_2} \quad (7.3)$$

where ε'_{eff} is the effective permittivity of the mixture, ε'_1 is the permittivity of the matrix, ε'_2 is the permittivity of the filler, and θ is the volume fraction.

From the mixing ratio of 1:1, θ is taken as 0.5, and ε'_1 , the permittivity of the cement, is taken as 2.5, and ε'_2 , the permittivity of lime, is taken as 1.9, the effective permittivity of the fabricated plaster is calculated to be 2.185. As the actual permittivity of the cement and lime used is unknown, this calculated static effective permittivity still provides an understanding on the effect on the S_{11} as permittivity generally increases with frequency. In addition, the values here are of static permittivity. From the previous chapter, section 6.4.3, with a lower permittivity, the S_{11} is also worsen. Therefore, embedding antenna in plaster has a worse effect than embedding antenna in cement paste, and it is not recommended.

7.4.2 Plaster only — effects on transmission coefficient

(a)

(b)

Fig. 7-11 Comparison between samples 23, 25, 03 and 13 after embedding (a) S_{11} , (b) S_{21}

Sample Ant23 and sample Ant03 (solid red and solid blue lines) in Fig. 7-11(b) show the S_{21} of the embedded metaresonator array in plaster only and cement paste only samples respectively. Both blue lines represent the cement paste samples while the red lines represent the plaster samples. For both pair of lines, the admixture added to the base material of plaster and cement paste contribute to the significant differences in the plot of S_{21} .

Focusing on the centre frequency at 2.45 GHz, Table 7-6 shows a summary of the S_{21} values at 2.45 GHz.

Table 7-6 Summary of S_{21} of plaster and cement paste samples

Sample Type	Sample No.	S_{21} (dB)		Change in S_{21} (dB)
		Before	After	
Plaster only	Ant23	-14.6	-25.0	10.4
Cement paste only	Ant03	-17.1	-25.9	8.8
Plaster with 4% iron (III) oxide	Ant25	-12.1	-34.5	22.4
Cement paste with 4% iron (III) oxide	Ant 13	-14.0	-20.1	6.1

From these values, the S_{21} decreases approximately by between 6 to 10 dB for all the samples except for plaster enhanced with 4% iron (III) oxide which shows a decrease of 22.4 dB. Analysing the plaster only and cement paste only samples, plaster only sample decreases the S_{21} by 10.4 dB while cement paste only sample decreases the S_{21} by 8.8 dB. This indicates that plaster material has a higher SE than cement paste. When 4% iron (III) oxide is added to both plaster and cement paste, the plaster with 4% iron (III) oxide sample decreases S_{21} by 22.4 dB while cement paste enhanced with 4% iron (III) oxide sample only decreases the S_{21} by 6.1 dB. Adding this admixture to plaster worsen the S_{21} than without any addition. On the other hand, when this admixture is added to cement paste, the S_{21} improves by 2.1 dB. This finding suggests that plaster is a worse material for embedded metaresonator array than cement paste.

Furthermore, adding 4% iron (III) oxide to the materials deteriorates the performance of plaster, while improving the performance of cement paste.

In this application of embedded antenna for buildings, the S_{21} is measured with the metaresonator array embedded in building materials (see Fig. 5-2(b)). The resulting observations of plaster only and cement paste in Fig. 7-11(b) can be understood by considering penetration depth in materials. Penetration depth is commonly defined as the decrease in power density of electromagnetic wave by e^{-1} of its initial value at the surface [203]. The relationship between a non-magnetic material ($\mu_r = 1$) and frequency [203][204] is given as

$$D_p = \frac{v\sqrt{2}}{\pi f} \left\{ \epsilon_r' \left(\sqrt{1 + \left(\frac{\epsilon_r''}{\epsilon_r'} \right)^2} \right) - 1 \right\}^{-1/2} \quad (7.4)$$

where f is the free space frequency, $v = \frac{1}{\sqrt{\epsilon_r \epsilon_0 \mu_0}} = \frac{c}{\sqrt{\epsilon_r}}$ is the propagation velocity in the material, ϵ_r' is the relative dielectric constant, and ϵ_r'' is the relative dielectric loss.

To illustrate, Fig. 7-12 is plotted to show the relationship of the penetration depth with increasing permittivity. From measurement data taken from [204], the ϵ_r' of concrete and plasterboard are 5.31 and 2.94 respectively, which signifies that D_p of plaster is higher than cement paste. However, it should be noted that differences exist between cement paste and concrete, and plaster and plasterboard. From [205], it is listed that the dielectric constant of plaster varies from 2.5 to 6.0 while from [206][207], the dielectric constant of cement paste varies from 2.5 to 7.0. In addition, from [208], the dielectric constant of Portland cement (which is used here) is 2.5 to 2.6, while that of plaster is 2.5 to 6.0. With these varying values, there are variations in the D_p calculated depending on the dielectric constants of the samples taken at the point in time. Having the results in Fig. 7-11(b), it suggests that the plaster used here has the same dielectric constant value as the cement paste within the WiFi spectrum. When 4% iron

(III) oxide is added to both materials, the D_p of iron (III) oxide enhanced plaster decreased significantly while the D_p of iron (III) oxide enhanced cement paste increased over the controls.

Fig. 7-12 Relationship between penetration depth and permittivity

7.4.3 Plaster layer on mortar — effects on reflection coefficient

Fig. 7-13 shows the S_{11} of the embedded metaresonators on cement paste with a layer of plaster. Samples Ant16 and Ant17 are the samples of metaresonator array on cement paste with a layer of plaster on it. Samples Ant20 and Ant21 are samples of metaresonator array on cement paste with a layer of 4% iron (III) oxide enhanced plaster.

Fig. 7-13(a) shows the comparison of samples Ant16 and Ant20. From 2.22 GHz to 2.6 GHz, the S_{11} values of both samples are very similar. Beyond 2.6 GHz, both samples exhibit similar trend of decreasing S_{11} values with a difference of about 2 dB. For samples Ant17 and Ant21 (see Fig. 7-13(b)), both S_{11} plots follow the same trend, with Ant21 having a 2 dB higher S_{11}

(a)

(b)

Fig. 7-13 S_{11} comparison after embedding between (a) Ant16 and Ant20 (b) Ant17 and Ant21

values in general. From 2.8 GHz and higher, both curves follow each other more closely. For all the four samples, the addition of plaster layer worsens the S_{11} at centre frequency while shifting the resonant frequency to be beyond 3 GHz. In this application, not only does the metaresonator array gets significantly mismatched, the addition of 4% iron (III) oxide enhanced plaster also does not have a significant positive impact on the S_{11} .

However, from sections 7.3.1 and 7.4.1, it is understood that plaster only sample worsen the S_{11} of embedded antenna by 6.5 dB while plaster with 4% iron oxide improves the S_{11} by 3 dB. Therefore, when plaster is added with 4% iron oxide, S_{11} of the embedded antenna changes significantly to be noticed. In this section however, by having a 1 cm layer of plaster on embedded antenna on mortar, there is no noticeable difference between the sample of plaster only on mortar and 4% iron oxide enhanced plaster on mortar.

This suggest that the layer of plaster applied is not thick enough to sufficiently show the impact of the addition of iron oxide. As the next section will explain, the penetration depth at this frequency is more than 1 cm for a wide range of permittivity.

7.4.4 Plaster layer on mortar – effects on transmission coefficient

In terms of S_{21} , from Fig. 7-14, the trend of the S_{21} curves are also very similar to the respective plaster only on cement and 4% iron (III) oxide enhanced plaster samples with the differences between the values within 2 dB to 5 dB. The exception is for samples Ant16 and Ant20 (see Fig. 7-14(a)), having a difference of 12 dB at 2.65 GHz, and more than 2 dB difference between 2.55 GHz to 2.7 GHz, and samples Ant17 and Ant21 (see Fig. 7-14(b)) of 10 dB at 2.85 GHz. Within the WiFi spectrum, there is little to no difference between the control samples and the samples with 4% iron (III) oxide enhanced plaster on mortar. This suggests that there is little effect between plaster and 4% iron (III) oxide enhanced plaster on mortar on embedded metaresonators.

From Fig. 7-12, the penetration depth is more than 10 cm even as the permittivity goes beyond 20. As the thickness of the plaster on mortar is only 10 cm, it is challenging to measure the

(a)

(b)

Fig. 7-14 S_{21} comparison after embedding between (a) Ant16 and Ant20 (b) Ant17 and Ant21

effects this layer of plaster has on the S_{21} . As the penetration depth is a measure of the intensity of the EM radiation that falls to about 37% from just beneath the surface, and that the permittivity of plaster is low, the amount of attenuation α given by [211] is

$$\alpha = \frac{\omega}{c\sqrt{2}} \sqrt{\epsilon_r' \mu_r'} \sqrt{\sqrt{1 + \tan^2 \delta_d} - 1} \quad (7.5)$$

where $\tan \delta_d$ is the dielectric media loss tangent. The unit of α is Np/m.

As $1 \text{ Np} = 20 \log_{10} e \approx 8.68589 \text{ dB}$, we can obtain α in terms of dB.

Plotting α assuming a loss tangent of 0.01 for concrete in Fig. 7-15, the amount of attenuation is about between 3 dB/m to 7 dB/m for permittivity of between 2 to 10.

Fig. 7-15 Relationship between attenuation and permittivity

With our plaster thickness of 1cm, the attenuation therefore will be between 0.3 to 0.7 dB for a material of effective permittivity of between 2 to 10. Therefore, with this small attenuation for the applied thickness of the plaster, it is logical that no apparent difference can be observed.

7.5 Concluding Remarks

The findings in Chapter 4 indicate that when admixture pigment of micro-sized iron (III) oxide is added to cement paste, the S_{11} performance of the embedded metaresonator array is improved, while the RF power transmission property of the material is increased at the WiFi spectrum. In Chapter 5, the results obtained further suggest that within the typical concentration of pigment used in the building industry, 4 weight percentage of micro-sized iron (III) oxide enhanced cement paste shows considerable improvement on the S_{11} of the embedded metaresonator array while significantly allowing more RF power to transmit into the material compared to the controls and other pigment concentration samples.

In this chapter, recognising the application of this research effort in buildings, an analysis of a common application of plaster on mortar is performed. In terms of S_{11} , embedding metaresonator array into plaster produces a negative effect on the antenna compared to cement paste, although the overall S_{11} is reduced and the bandwidth increased. When the plaster is enhanced with 4% iron (III) oxide, the enhanced plaster did not significantly change the resonant frequency or the minimum magnitude of S_{11} , although the bandwidth is increased. This suggests that adding 4% iron (III) oxide to plaster can increase the bandwidth while maintaining the resonant frequency and minimum S_{11} value. When plaster is applied as a layer on mortar with the metaresonator array embedded between the layer of plaster and mortar, all the three samples showed a significant distortion in the S_{11} plots. The resulting S_{11} values are also not as good as the previous results with general values at approximately -6 dB. Even when the plaster layer is enhanced with 4% iron (III) oxide, the results obtained are similar to the previous three samples without any enhancement. This finding shows that embedding the metaresonator arrays in between plaster (enhanced or not) and mortar significantly affected their performance to the point of unusable. These results show that while plaster enhanced with 4% iron (III) oxide increases the bandwidth and maintain the resonant frequency of the embedded metaresonator array, embedding these antennas between a layer of plaster and mortar is not appropriate for embedded WSN.

In terms of S_{21} , the results of plaster only sample is similar to cement paste only samples where the S_{21} decreases, indicating plaster and cement paste reduces the amount of RF power transmission into the material. However, unlike the cement paste enhanced with 4% micro-sized iron (III) oxide, plaster enhanced similarly produces a significant negative effect on S_{21} . From the results, adding iron (III) oxide decreases the maximum S_{21} by 20 dB while plaster only sample only shows a decrease of 11 dB. Compared to cement paste samples, cement paste only sample decreases the maximum S_{21} by 9 dB while the cement paste enhanced with 4% micro-sized iron (III) oxide gives a 6 dB decrease. These finding indicates that both plaster and enhanced plaster are not suitable materials for the embedded metaresonator array. When plaster is layered on mortar with the array embedded between the two materials, no significant changes can be observed. Between plaster layer on mortar and iron (III) oxide enhanced plaster layer on mortar, no substantial difference is noticed either. This finding suggests that embedding the metaresonator array in between a layer of plaster and mortar does not produce improvement on either the impedance matching of the antenna or RF power transmission property of the material, irregardless the layer of plaster is enhanced with iron (III) oxide or not.

Chapter 8

CONCLUSION

8.1 Conclusion

It is a global trend that cities are building new or retrofitting existing buildings with advanced technologies to make them smarter, more sustainable and green. To enable this movement, deploying WSN in indoor spaces to gather and transmit useful data about the environment is recognised by many governments as part of the green building standards worldwide. In the future, embedding these sensors in prefabricated building structures is a possibility that can promote non-invasive monitoring of the building throughout its entire life cycle. To realise this goal, embedded WSN needs a source that can provide power for a prolonged duration, and an effective way to transmit data out of the building materials. External RF sources is considered a promising method to provide such continuous electrical power within the built environment, however building materials have long been known to have high SE preventing good wireless propagation in the indoor environment. The contributions in this thesis thus focus on designing the antenna in the embedded WSN system without battery to receive external RF energy sources and wireless communications. In addition, by investigating common admixture pigments used in the building industry, a suitable iron-based oxide is identified, along with the size and concentration, as well as a method of applying this admixture with embedded metaresonator arrays in building materials.

The key properties of the antenna component in embedded WSN systems suitable for receiving RF energy from external sources and wireless communications includes planar in design, scalable, low cost, and adjustable to suit different deployment set-up in buildings. To fulfil these criteria, a study of the concepts of metamaterial and electrically small antennas was done to determine that when a type of antenna is limited to a maximum dimension of one-tenth the wavelength, these two concepts can be combined to form antenna arrays with small element separations. However, to begin the element design of a gigahertz ESA of 0.1λ is challenging with the conditions mentioned for use in buildings.

To realise such an antenna, the concept of SRR is combined with ESA to form metaresonators. A gigahertz electrically small metaresonator is designed using low cost materials and standard PCB processes in Chapter 3. In addition, this metaresonator is planar and can also function as an element in an array. The main advantageous feature of this antenna is its stubbed design. The length of the stubs can be adjusted to tune the antenna to the desired resonating frequency. While the antenna is well matched with a S_{11} of -25 dB, the efficiency of -9 dB and gain of -16 dB is within expectations of an ESA.

Another key feature of this electrically small metaresonator is its geometry that allows it to tessellate well over a surface. The second contribution in Chapter 3 shows that by selecting the maximum element size to be one-tenth the wavelength, the concepts of metamaterials and antenna array can be combined to form an antenna array such that the elements can be arranged close together. The element separation optimised in this thesis are 0.000724λ and 0.0165λ for the two different configurations simulated, which is much less than other such arrays in the literature of 0.161λ [167] and 0.202λ [168]. Using one of the array configurations, the S_{11} obtained for type B configuration is -25 dB with an efficiency of -6 dB and a gain of -4 dB. It is also further analysed that by feeding this array at different feed points, the properties of the array such as the radiation pattern and gain can be altered, making it suitable for building applications.

The design of the antenna allows flexible implementation of the antenna to either function as a single antenna or scaled into an array. The ability to scale up into an array of different sizes

allows the antenna to suit different surfaces in a building. At the same time, the properties of the array can also be modified to suit different application needs in the building. Being planar and low cost, this design can be mass deployed in a building to line the surfaces such as façade, walls, ceilings, floors and partitions. Therefore, this array design meets the objective of having a scalable antenna that can function both as a wireless communications antenna, and a RF EH antenna in the buildings for embedded WSN.

The next contribution in this thesis is addressing the poor wireless propagation in buildings due to high SE of building materials, specifically in the transmission of RF power in and out of building materials for embedded WSN. Admixture pigments are commonly used in the building industry for colouring concrete or mortar. In Chapter 4, cement paste enhanced with two different iron-based oxide admixtures in two different sizes are examined. From this finding, it is shown that both micro-sized particles of magnetite and iron (III) oxide 1) improve the impedance matching of the embedded arrays, 2) increase the bandwidth of the array, and 3) maintain the overall shape of the S_{11} with a lesser shift in resonant frequency than cement paste alone. Further analysis in S_{21} shows that adding micro-sized iron (III) oxide into cement paste improves the RF power transmission over cement paste only sample by as much as 10 dB.

In the building industry, the allowed weight percentage of admixture pigment used is regulated to be less than 10%. Typically, these pigments are mixed with cement with percentage of less than 5% due to diminishing colour changes from 6% and above. In Chapter 5, cement paste samples enhanced with 0% to 4% of micro-sized iron-based oxide are prepared to determine the concentration of this admixture pigment that can improve both the performance of the embedded metaresonator array and the RF transmission property of the building material. The results in this chapter find that 4% micro-sized iron (III) oxide enhanced sample improves the S_{11} of the embedded metaresonator by as much as 20 dB and a S_{21} improvement of up to 7.29 dB over the cement paste only sample.

The combined contributions in Chapters 4 and 5 show that by enhancing cement paste with 4% micro-sized iron (III) oxide particles, the S_{11} and bandwidth of the embedded metaresonator array are improved, while the amount of RF power transmission in and out of the building

material is increased. This meets the objective of improving building materials for RF EH and wireless communications for embedded WSN.

In Chapter 6, further analysis on application of plaster is done to develop an understanding of certain applications in buildings. It is found that while plaster enhanced with 4% iron (III) oxide alone can increase the bandwidth of the embedded metaresonator array while maintaining the minimum S_{11} value and resonant frequency, there is no improvement in the RF power transmission property of the material. In fact, the results obtained shows that plaster samples are the same or worse than cement paste samples. Furthermore, the findings here indicate that embedding the metaresonator array in between a layer of plaster (enhanced or not) and mortar significantly deteriorates the S_{11} to render it unsuitable for wireless applications. Analysing S_{21} , it is observed that plaster enhanced with iron (III) oxide layered on mortar also does not significantly change the RF transmission property of the material. The contribution here thus shows that embedding the metaresonator array in plaster or plaster layer on mortar is not suitable for the embedded WSN.

8.2 Recommendation for future work

This thesis focused on two important components in embedded WSN for green buildings, namely the antenna and the building materials. The design and investigation presented in this work contribute to the larger goal of enabling buildings to be smarter and greener. There are still some unanswered questions and a few related research directions worth pursuing:

- (i) *Scalable antenna array for RF EH and wireless communication:* In Chapter 3, two possible configurations are presented. While type B configuration can function well, other possible configurations such as having different orientation of elements, different number of elements, different length of stubs for each element can be considered. Other designs like adding resonators to improve the gain and efficiency of the array can also be studied. In addition, a rectifier circuit can be connected to this array to investigate the efficiencies and amount of RF power that can be harvested for different configurations and sizes.

- (ii) *Admixtures in cement paste:* Only iron-based oxide admixture pigments are tested in Chapter 4 and 5. Other admixtures such as copper slag, titanium dioxide and calcium nitrate are some of the other common admixtures used in the building industry. These materials are also interesting admixtures for investigations. Also, more samples can also be prepared to study the variation in samples to establish an understanding in other areas such as period of curing over extended periods and subjection to temperature cycle. Other types of studies such as embedding different antennas and arrays in enhanced building materials, investigations into effects on the antennas and materials are also worthwhile experiments to carry out. Larger samples can also be fabricated to better represent the large surfaces present in buildings.
- (iii) *Measurement and modelling of enhanced cement pastes:* Thus far, the contributions in this thesis have been on the RF measurement of the metaresonator arrays and materials. In the future, bigger samples can be fabricated to collect S_{12} and S_{21} values of the enhanced materials in addition to those measured here. With these values, Nicholson-Ross-Weir mathematical model can be applied to calculate the permittivity and permeability of the materials. These two values can then be used in the antenna simulation software to better design antennas suitable for embedding in such materials.

Appendix A

Comparison of Sample Between 3 Days and 30 Days

The comparison between the same sample of cement paste only (control) between 3 days and 30 days are shown in Fig. A-1 and Fig. A-2. Fig. A-1 shows that the S_{11} behaviour does not deviate by more than 2 dB.

In terms of S_{21} , Fig. A-2 also shows that the S_{21} behaviours are consistent between 3 days and 30 days. These results show that the behaviours of S_{11} and S_{21} of the antenna embedded in the samples of 3 days old and 30 days old do not deviate significantly.

It is also worth noting that, using the same “dry” curing conditions described in [202], the observations are similar to the results in [202] where the values of $|r|$ rapidly decreases during the first few days.

Fig. A-1 Comparison of S_{11} of cement paste only (control) between 3 days and 30 days from 2 to 3 GHz

Fig. A-2 Comparison of S_{21} of cement paste only (control) between 3 days and 30 days from 2 to 3 GHz

Appendix B

Measured S_{11} and S_{21} of Metaresonator Arrays

B.1 0% iron (III) oxide in mortar (control)

(a)

(b)

 Fig. B-1 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 1 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-2 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 2 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-3 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 3 (a) before embedding (b) after embedding

B.2 2% iron (III) oxide in mortar

(a)

(b)

Fig. B-4 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 4 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-5 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 5 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-6 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 6 (a) before embedding (b) after embedding

B.3 1% iron (III) oxide in mortar

(a)

(b)

Fig. B-7 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 7 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-8 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 8 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-9 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 9 (a) before embedding (b) after embedding

B.4 3% iron (III) oxide in mortar

(a)

(b)

Fig. B-10 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 10 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-11 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 11 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-12 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 12 (a) before embedding (b) after embedding

B.5 4% iron (III) oxide in mortar

(a)

(b)

Fig. B-13 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 13 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-14 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 14 (a) before embedding (b) after embedding

(a)

(b)

Fig. B-15 S_{11} (Blue solid line) and S_{21} (Red solid line) of Sample 15 (a) before embedding (b) after embedding

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